J. R. HALSALL

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Commentary

THAT all is not well with the present economy of this country has been made very apparent to all of us during the last few months. The reasons for our economic difficulties are many but one of the key issues is that the rate of the country's exports is not at a high enough level.

We have to export to live because this is a country with few natural resources and in the post-war years our income from overseas investments has disappeared.

The new Minister of Technology, Mr. Frank Cousins, in his recent address to the Grand Council of the Federation of British Industries viewed our present position with concern and in order to survive "we must live", he said, "by the products of our scientific and technological wits.

"The key to our present position must be our engineering exports," he continued, "and here we have gone down during the last nine years from 24 per cent of the world's markets to 18 per cent, and the greatest decline seems to be sharpest in the most technologically advanced products."

And why are we failing to sell our most technologically advanced products to the rest of the world?

Here again many reasons have been put forward—we have not, it is said, modernized our industry to the extent we should have done, we have not spent a sufficient amount of money on research and development and so on, and this is why our economy is not expanding fast enough and why our exports are failing.

Another reason which has been submitted on several occasions recently is the shortage of adequately qualified scientists and engineers.

This matter was dealt with recently by Col. G. W. Raby who, in his Presidential address to the Institution of Electronic and Radio Engineers, stated that according to figures produced by the Council of Engineering Institutions, the joint Corporate Membership of the thirteen Chartered Institutions is less than 130 000—and this includes membership abroad.

Even if this figure is doubled to include qualified engineers who are not members of a learned Institution —and this he considered a gross overestimate—it means that every qualified engineer in Great Britain carries a direct burden and responsibility for the well being of 200 of his countrymen every day he goes to work.

And then we express surprise, Col. Raby continued that the growth rate of our national economy is falling short of the predicted four per cent.

This shortage of qualified engineers and scientists is one of the problems which have been studied by the Advisory Council on Scientific Policy whose final annual Report has recently been published.

It will be recalled that the Council was established in 1947 "to advise the Minister of Science in the exercise of his responsibilities in the formulation and execution of Government scientific policy" but now the Council is coming to an end and due to the present reorganization of governmental civil service it will be replaced by a new Council for Scientific Policy.

Its main issue, and the most enduring issue which has exercised the council during its 17 years has been the question of the scale and balance of the national civil scientific effort, and the council admits that it has done no more than achieve a measure of the magnitude of the problem and on the great responsibilities which lie ahead.

After dealing at considerable length with the unsatisfactory amount of money spent on research and development the council goes on to look at this problem of the shortage of properly trained scientists and engineers in the industrial field in the hope that an adequate supply will solve all our economic difficulties.

"If our industry is to succeed," the Council's report states "it must be technologically in the forefront and it can only be so if it is able to recruit adequate numbers of suitably trained scientists and engineers; and industry must get its fair share of men of the highest quality."

The problem, it seems, is not that there is an overwhelming shortage of scientists—on the contrary the Council admits in its report that the output of scientists is five times what it was when the Council came into being 17 years ago—but far too many of these scientists are being attracted by the glamour of pure science.

"Already much concern has been expressed about the tendency for too high a proportion of our best brains to go into pure science than into applied science and engineering," says the Council.

The blame it appears rests with our universities where engineering has attracted too little of the best talent among the student population.

"The fact is," the Council states, "that the importance of academic research in chemistry, in physics and even in biology, and the ready utilization of the results of such research by industry can be easily observed and is publicly acknowledged, whereas the application of research in engineering and technological subjects expressed in terms of approved techniques is frequently overlooked and passes unacclaimed. This fact, combined with the greater academic prestige of the pure sciences at the universities, has certainly an effect upon the attractiveness of these subjects as compared with engineering."

So the solution is apparent. We have to attract more of our brighter people to engineering and then all our problems will have been solved. We have, of course, to find sufficient money "to enable our scientists fully to display their ability and provide the essential background to our technological effort in industry as well as producing the trained manpower which industry requires".

An X-Band Spectrum Analyser and Sweep Generator

By R. W. Eldred*, A.M.I.E.E.

This article describes briefly a combined X-band spectrum analyser and sweep generator. The sweep generator arises from the particular method employed to ensure constant sensitivity in the spectrum analyser over the electronic sweep range. A method of using an internal absorption-type wavemeter in an a.f.c. loop to stabilize the local oscillator frequency under 'manual sweep' conditions is also described.

(Voir page 64 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 71)

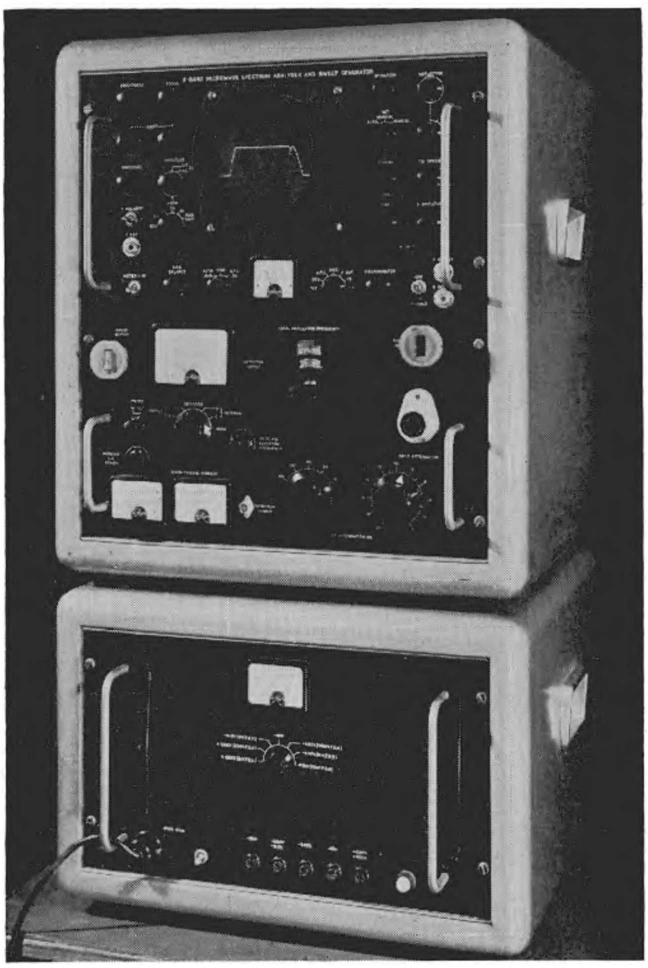
THE instrument described in this article was designed to have the following features:

(1) Accurate spectrum measurement over a wide electronic tuning range.
(2) Good resolution.
(3) Low susceptibility to pick-up from external sources.
(4) Good local-oscillator stability.
(5) Accurate measurement of frequency, without ambiguity.
(6) Either automatic or manual sweep.
(7) Large screen display.
(8) No time-lag on any of the controls.

The methods used to achieve these objectives lead to the inclusion of two additional features not normally associated with a spectrum analyser viz.:

(9) A source of constant power output for use as a sweep generator.
(10) Stabilization of the local-oscillator frequency at any point in the tuning range of the instrument.

The complete equipment



Items (1) to (8) are required to simplify the analysis of complex spectra obtained in frequency-modulated c.w. systems.

If a reflex klystron local-oscillator is used, the sensitivity of the instrument will normally decrease considerably at the extremes of the electronic tuning range due to the parabolic power-output versus reflector-voltage characteristic. This effect has been eliminated in the present design by using a local-oscillator klystron with a higher output power than would normally be used. The local-oscillator power at the microwave mixer is then stabilized at the required level by means of an automatic power-level controller. The oscilloscope trace in the photograph shows the constant local-oscillator power characteristic, with the wavemeter dip on the extreme right.

A fraction of the output from the power-level controller is coupled to the waveguide output on the left of the lower panel. This is normally terminated with a matched load during spectrum analysis. The electronic sweep width is approximately 30Mc/s with a constant output of 1mW. This can be increased to 60Mc/s at certain parts of the mechanical tuning range by reducing the sweep output power to about 0·5mW.

The internal wavemeter is included in the local-oscillator circuit rather than in the signal input circuit, thus ensuring that the frequency to which the instrument is tuned may be accurately determined. Since the power level at the input to the wavemeter is stabilized, the wavemeter can be used in an a.f.c. loop to reduce frequency drift when the instrument is manually tuned to a particular frequency of interest.

The mechanical tuning range of the instrument is from 8·5Gc/s to 9·5Gc/s.

General Description

A simplified block diagram of the instrument is shown in Fig. 1.

The output from the local-oscillator is passed through an isolator to an electronically controlled attenuator, both of these components being microwave ferrite devices employing the Faraday rotation effect. The output from the variable attenuator is divided into two paths, one of which is again divided in a second short-slot hybrid.

Under normal operating conditions a power level of 2mW is fed to the

* *The Marconi Co. Ltd.*

microwave mixer via an absorption wavemeter. The microwave signal input is fed to the balanced mixer via a calibrated vane-type attenuator.

A power level of 1mW is fed to the local oscillator power-level control loop, and a further 1mW is fed to the sweep generator output. A microwave vane-type attenuator determines the amount of power fed to the crystal detector in the power-level control loop, and this in turn determines the control current in the electronically-controlled attenuator. The power level at the sweep generator output can therefore be varied by means of the manually operated attenuator in the microwave circuit at the input to the control loop.

All the supplies for the local oscillator are stabilized, including the d.c. heater supply. The reflector-voltage stabilizer is used as a virtual earth amplifier in the electronic sweep circuit, the manual sweep circuit, and the local-oscillator frequency stabilization circuit.

To stabilize the frequency of the local-oscillator the manual tuning control is set to the required frequency and the wavemeter is then adjusted to a position which gives approximately half the maximum absorption, i.e. the required frequency is situated on one side of the wavemeter response. Under these conditions the wavemeter converts local-oscillator frequency variations into amplitude variations at the microwave mixer input.

A d.c. voltage proportional to crystal current in one of the mixer crystals is amplified in a transistor amplifier and applied to a d.c. balancing network. The output from this network can be brought to zero at the required operating point. Any deviation of the local-oscillator from the required frequency produces an output from the 'discriminator' circuit, which is further amplified and applied to the reflector voltage stabilizer for frequency correction.

When the instrument is operating in its normal mode as a spectrum analyser, microwave input signals are converted to the intermediate frequency and passed through a precision step attenuator and a high gain transistor amplifier. The rectified i.f. output is then amplified in a balanced d.c. amplifier and passed to the Y-deflexion amplifier.

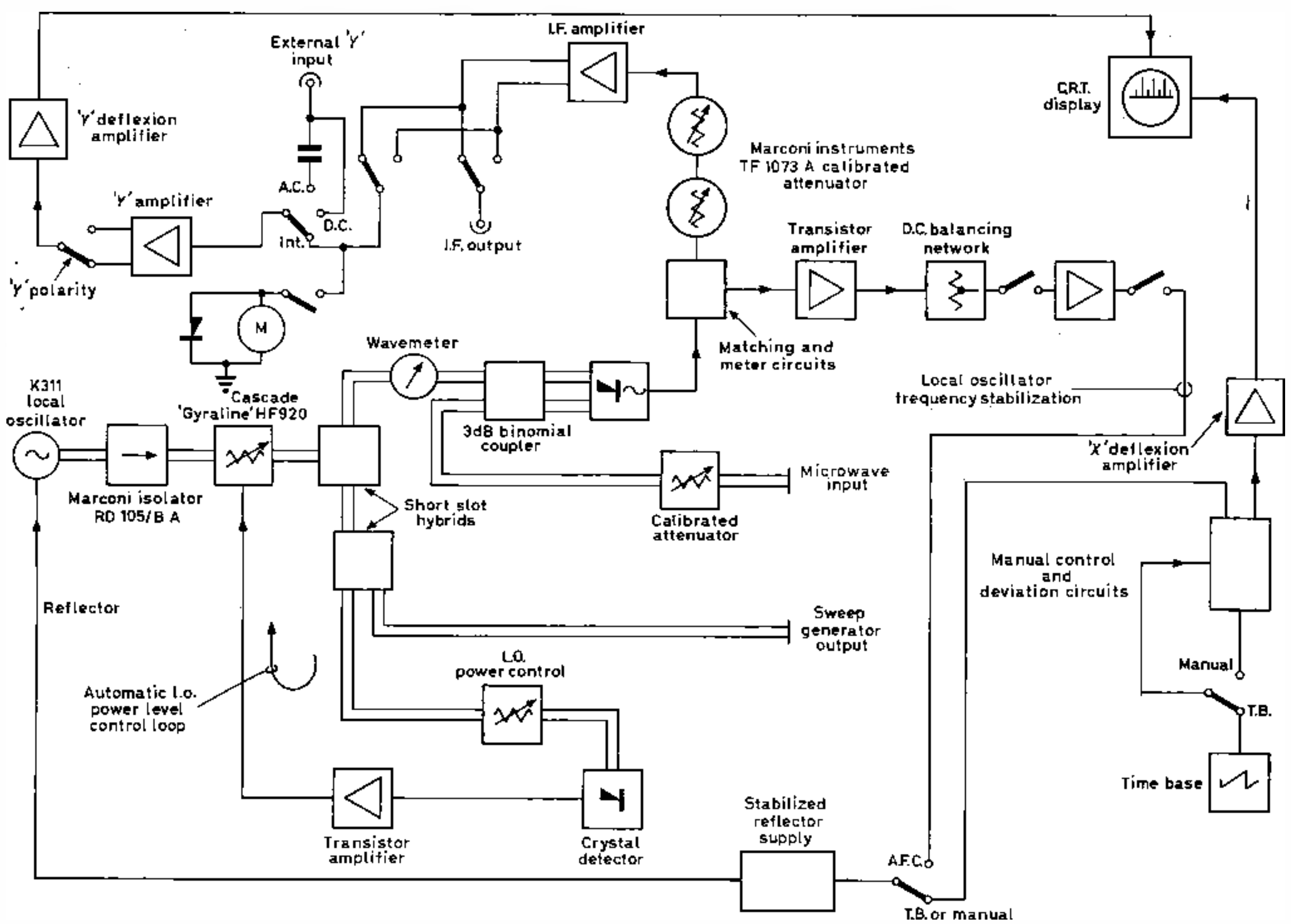
The power supplies are housed in a separate cabinet to reduce supply-frequency pick-up and also to reduce heat dissipation in the spectrum analyser cabinet. The latter consists of two units: the waveguide unit at the bottom of the cabinet, and the oscilloscope and control unit at the top. Transistor amplifier circuits for the local-oscillator power control loop and the frequency stabilization loop are housed in the waveguide unit. The transistor i.f. amplifier, which is completely screened, is also situated in the waveguide unit close to the microwave mixer.

Time-Base Circuits

The time-base circuits are shown in Fig. 2. Sweep speeds between approximately 0·5 and 10c/s are available. By using d.c. coupled circuits throughout, very low time-base speeds can be used without distortion of the sweep or time-lag on the controls.

The frequency deviation control (R_3) is included in a balanced d.c. circuit so that the deviation can be varied

Fig. 1. Simplified diagram of the equipment



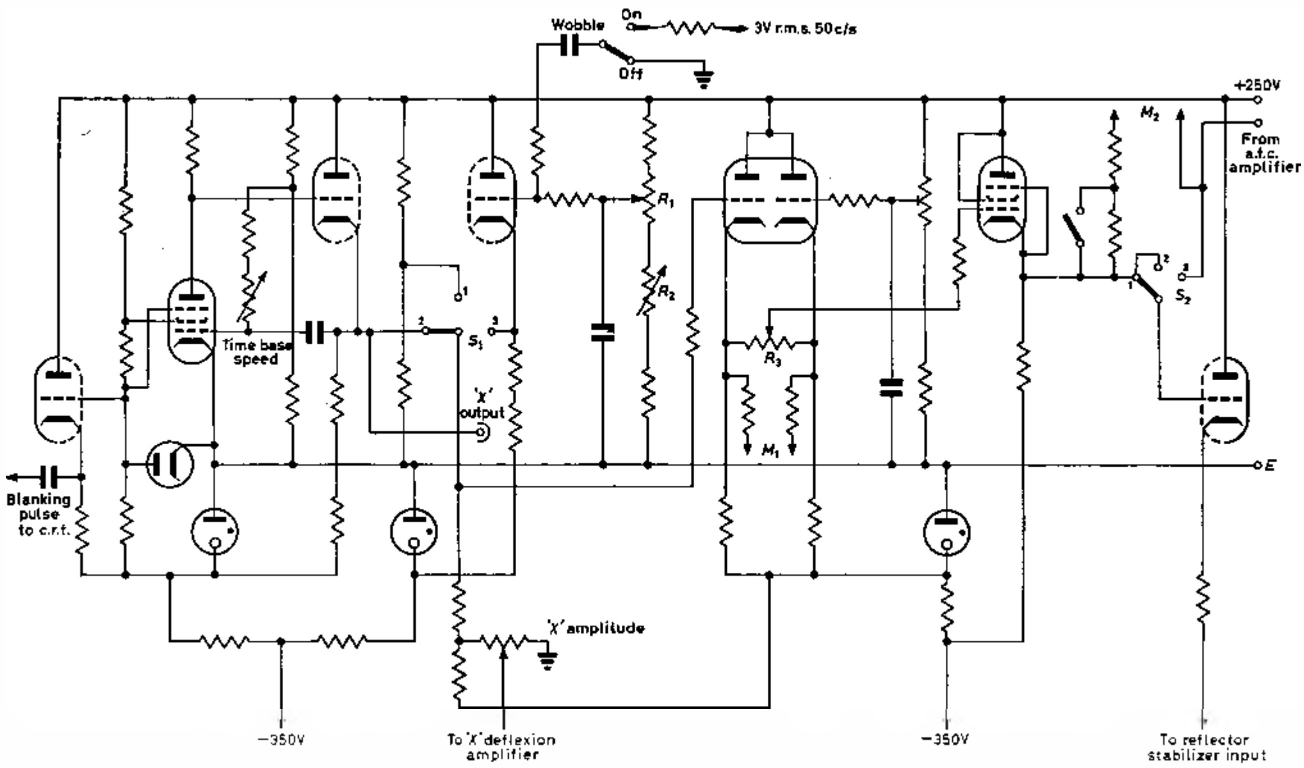
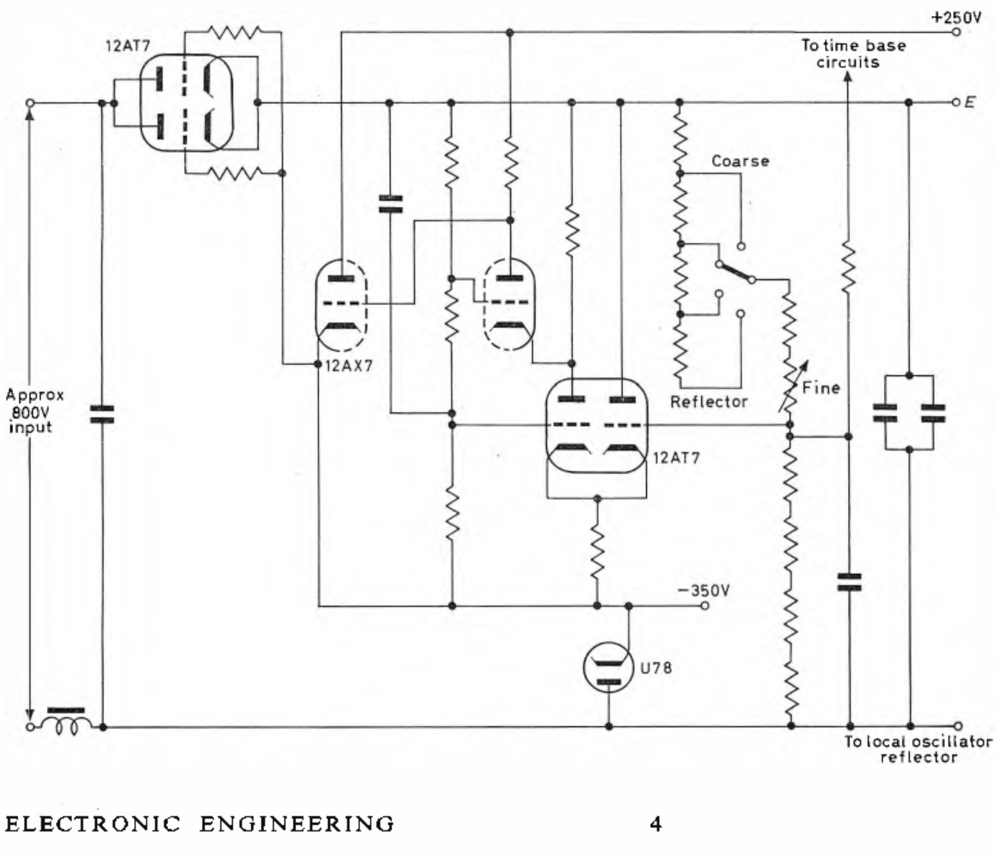


Fig. 2. Time-base circuits

without changing the mean frequency. When switch S_1 is at position 2, the time-base output is fed to the X-deflexion amplifier via the X-amplitude control, and also to the reflector voltage stabilizer via the deviation control circuit. Thus the klystron reflector voltage and the oscilloscope deflexion are swept in synchronism.

Setting switch S_2 to position 3 allows the klystron reflector voltage and the X-deflexion circuit to be swept manually, by means of potentiometer R_1, over exactly the same range as for automatic (repetitive) sweep. When manual sweep is in use a small frequency 'wobble' at the mains supply frequency can be switched on. This allows any part of a complex spectrum to be 'illuminated' for detailed investigation.

Fig. 3. Klystron reflector-voltage stabilizer



A complete display of the klystron power characteristic over the whole tuning range (8·5Gc/s to 9·5Gc/s) requires a maximum available reflector voltage sweep of approximately 90V. The most economical method of sweeping the reflector voltage is to use the reflector-voltage stabilizer as a virtual-earth amplifier. This arrangement is shown in Fig. 3.

A compromise design is necessary since the shunt capacitance in the stabilizer feedback circuit should have a large value to provide maximum loop-gain at ripple frequencies, but a low value to allow a reasonably fast sweep. If the sweep speed is too fast the stabilizer will be overloaded during the fly-back period, and distortion of the forward-sweep will also occur.

The Microwave Circuit

The isolator in the klystron output circuit is necessary to prevent the variation in input impedance of the electronically-controlled attenuator from pulling the klystron frequency. An isolation of approximately 40dB is provided at 9Gc/s, falling to 20dB at 8·5Gc/s and 15dB at 9·5Gc/s.

Under normal operating conditions the output from the electronic attenuator is 4mW. If the maximum output power from the klystron is 50mW, an electronic control range of 11dB is required as the minimum. To provide a wide electronic frequency sweep at low power a 20dB range of control is desirable. The attenuator used has a 'zero-current' attenuation of 3dB, allowing the output to be decreased *or* increased electronically.

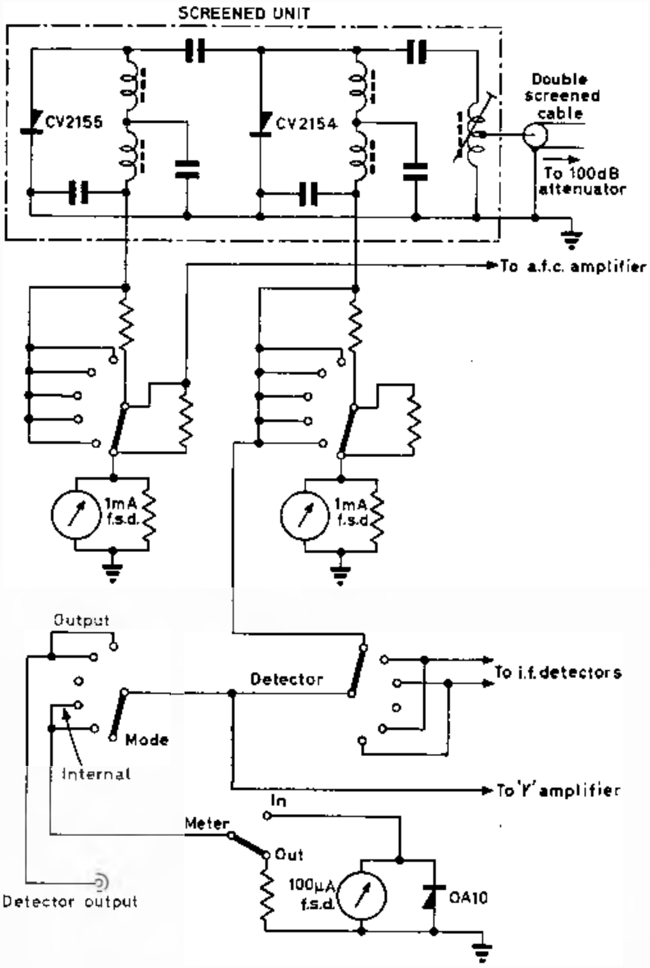


Fig. 4. Microwave mixer circuits

The loaded Q factor of the absorption-type wavemeter is approximately 10 000.

The attenuator in the signal input circuit is of the moving-vane type with a micrometer screw adjustment. A maximum attenuation of at least 40dB is provided.

Fig. 4 shows the microwave mixer circuit and the i.f. detector switching circuits. The crystal current meters are shunted to read approximately 2mW f.s.d., the meter switching being included on the i.f. detector switch. Additional series resistance is included in the meter circuits when the klystron power characteristic is displayed.

Local Oscillator Power Control

A d.c. voltage proportional to the current in the micro wave crystal detector (Fig. 1) is amplified in a transisto d.c. amplifier (Fig. 5). As the microwave power leve increases from zero the output from the second OC201 ir Fig. 5 approaches 3V. When the output just exceeds 3\ current flows through the control winding of the variabl attenuator, preventing further increase of the local-oscil lator power.

A current of 0·5mA flows through the crystal whe the microwave power is zero. The current increases t approximately 1mA under normal operating conditions when the current attenuation between the input to th attenuator and the input to the transistor amplifier is 40dB This means that drift in the input circuits of the d.c amplifier is very important and should be reduced to : minimum.

The response time of the power-level control loop i determined by the d.c. gain of the complete loop and th open-loop bandwidth. During manual sweep the respons time of the system is not important since it can easily b made much smaller than that of the operator. However with a repetitive sweep the closed-loop time-constant mu: be small to ensure a short settling-time once the micro wave power exceeds the 'delay' in the loop.

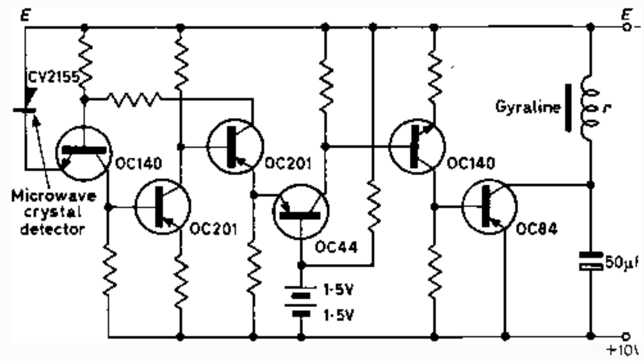


Fig. 5. Automatic local oscillator power control

The final design of the power-level control loop was : compromise between loop stability and time-base speed In the photograph (page 2) the time-base frequenc was approximately 10c/s; it will be observed that th settling time was quite adequate for this frequency.

If the response time of the loop is too long the shap of the power characteristic is affected by the 'clippin level'. This is due to the pulse of microwave power durin the flyback time stimulating a response in the power con trol loop. The effect is particularly noticeable at high time base frequencies when the flyback time is comparable wit the closed-loop response time. In this condition the control current in the attenuator is not zero at the start of th forward sweep and the left-hand edge of the power charac teristic is distorted (when sweeping from left to right).

Relatively high-frequency transistors should be used i the d.c. amplifier to ensure a good margin of stability i the loop. Most of the transistors in Fig. 5 are used in th grounded-emitter configuration, which leads to reduce bandwidth per stage compared with the grounded-base con dition. Excluding the main time-constant of the loop, th principal bandwidth limiting components are the OC8 power transistor and the current-controlled attenuator The latter has a modulation bandwidth of 4kc/s.

As the klystron reflector voltage is swept across th required range, the current flow through the attenuato varies between zero and a maximum of approximatel

0mA. Maximum attenuation, and therefore maximum ontrol current, is required when the klystron output power ; maximum. In the region of maximum attenuation the verall current gain of the d.c. amplifier is 94dB, giving a oop gain of 54dB.

The open-loop 3dB bandwidth is approximately 100c/s, eing determined by the resistance of the attenuator conrol winding and the 50μF capacitor. The main bandwidthimiting circuit has been placed at the terminals of the ttenuator, rather than at an earlier stage in the amplifier, ɔ reduce noise modulation of the microwave signal.

With the circuit shown in Fig. 5, a 1dB reduction in ɔcal-oscillator power at the microwave mixer was neasured as the temperature of the d.c. amplifier was ncreased from 20°C to 40C°. This performance is adeuate for normal laboratory use.

Assuming wideband microwave components and a erfectly matched microwave circuit, the constancy of the ɔcal-oscillator power would be determined by the characeristics of the 'delay' cells and the loop gain only. In ractice, a tilt occurs on the power characteristic at certain requencies due to the varying conditions of match in he waveguide circuit. As the local-oscillator is tuned nechanically over the range 8·5Mc/s to 9·5Mc/s the slope f the 'flat' portion of the power characteristic varies rom positive through zero to negative and back again everal times.

The degree of tilt with the microwave components used oes not exceed ± 10 per cent and this does not impair ne accuracy of the spectrum analyser. An external matchng section can be used to correct the tilt when the instrunent is used as a sweep generator if desired, but this is ot normally necessary.

The sweep generator facility provided in this instrunent is particularly useful for tuning klystron amplifiers nd narrow-band microwave filters. When used for this urpose the input to the balanced Y-amplifier is accessible rom the front panel. The oscilloscope sensitivity can be aried between 10mV/cm and 1V/cm. A Y-polarity eversal switch is also provided.

ocal Oscillator Frequency Stabilization

When the klystron has been manually tuned to a frequency at which stabilization is required, the wavemeter ; tuned to give a 10 per cent reduction in local oscillator ower. For small frequency variations the mixer crystal urrent is then approximately proportional to frequency nd the output from the a.f.c. amplifier is approximately roportional to frequency.

When the local-oscillator has been tuned to the required requency and the wavemeter correctly adjusted, the negaive bias voltage in the 'discriminator' circuit (Fig. 6) is djusted to be equal in amplitude to the positive output oltage from the transistor d.c. amplifier. The output of he second d.c. amplifier in Fig. 1 is then adjusted to be qual to the output from the time-base circuits (at position , of S_2 in Fig. 2) and switch S_2 is moved to position 3 to lose the a.f.c. loop.

The voltage loss between the klystron reflector and the nicrowave mixer output is 34dB, and a 6dB loss occurs n the 'discriminator' circuit. The d.c. gain of the tranistor amplifier is 34dB, so the net loss between the klystron eflector and the 'discriminator' output is 6dB.

The frequency drift of the klystron without stabilization but with all supplies stabilized) did not exceed 100kc/s ver reasonable periods of time. With an i.f. bandwidth of 5kc/s ($\pm$12·5kc/s) it is desirable to reduce the drift below kc/s. This requires a stabilization loop-gain of at least

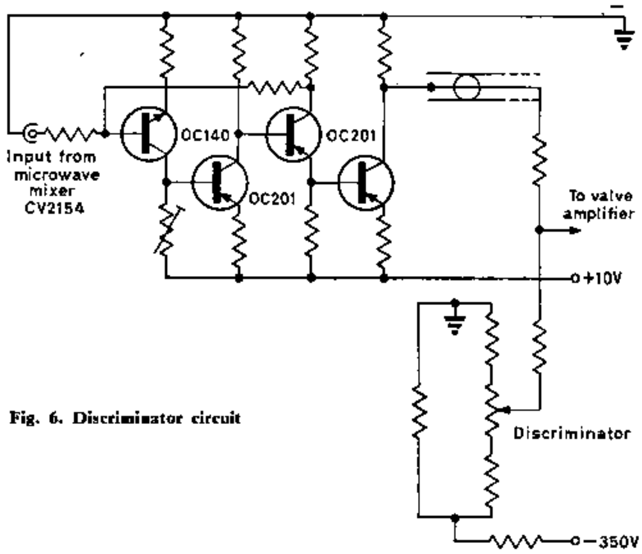


Fig. 6. Discriminator circuit

40dB. The total loop-gain provided is 52dB to allow for deterioration. The open-loop bandwidth is approximately 6c/s and this gives a closed-loop bandwidth (3dB) of 2 400c/s.

When the local oscillator frequency is stabilized a small range of frequency adjustment can be made by tuning the wavemeter. The output from the i.f. detector can then be measured on a 100μA meter.

The output from the sweep generator may also be used as a frequency-stable X-band signal. Under these conditions the instrument has been used as a stable localoscillator when examining frequency drift in other microwave oscillators due to power supply variations, etc.

The transistors in the d.c. amplifier are mounted in a large heat sink which is bolted to the chassis of the waveguide unit to provide a long thermal time-constant.

I.F. Circuits

A transistor double superhet' i.f. circuit is used, the first i.f. being 22·5Mc/s and the second i.f. 1·5Mc/s. Diode detectors follow the final and penultimate stages. The overall i.f. bandwidths are 25kc/s and 30kc/s respectively. Receiver noise is clearly displayed on the oscilloscope when the total i.f. power gain of 107dB is used, and a 10 per cent deflexion is obtained on the detector-output meter.

A 100dB step attenuator, adjustable in 1dB increments, is used at the i.f. amplifier input for accurate spectrum measurement. The dynamic range of the receiver is 60dB, although the dynamic range as determined by receiver noise is about 90dB. The dynamic range is limited by the second mixer in the present design.

Oscilloscope Design

The oscilloscope is designed around a flat-faced longpersistence 6in diameter c.r.t. Both the X and Y deflexion amplifiers and the shift controls are d.c. coupled.

When used for spectrum analysis the Y deflexion circuits are d.c. coupled from the i.f. detector output right through to the Y-deflexion plates of the c.r.t. The oscilloscope may be used either as an a.c. or d.c. oscilloscope in its own right. The bandwidth of the Y-amplifier is 300kc/s at 1V/cm sensitivity, falling to 3kc/s at 10mV/cm.

Acknowledgments

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Experiments in M.H.D. Power Generation

By G. J. Womack*, M.Sc.Tech., Ph.D., A.M.Inst.F.

This article reports results obtained with a small gaseous propane-oxygen combustion driven m.h.d. generator. A comparatively large permanent magnet was used to produce a constant flux density of 9 800 gauss. The combustion products were seeded with up to approximately 1·6 per cent by weight potassium. Experiments were carried out with both hot carbon and water-cooled copper electrodes.

Electrical conductivity determinations of the working fluid have been made, using both types of electrodes, by measuring the I-V characteristics under induced and applied electric field conditions. To assist the interpretation of these results a double beam sodium line reversal apparatus was built to measure the gas temperature. Good agreement between the measured and theoretically calculated electronic conductivity was obtained for both hot and cold electrodes. For the cold electrodes the mode of the conductivity was electronic and not ionic, therefore the cold cathode must be electron emitting. The exact mechanism of this emission is not fully understood, but is probably due to carbon deposition on the electrode surface. This carbon is probably amorphous and of high porosity and low thermal conductivity. Therefore, this layer will approach the stagnation conditions of the gas stream which would be sufficient to give high thermionic emission.

From the electrical conductivity measurements the collision cross-sections have been determined and compared to the theoretically predicted values calculated by Frost's method[14]. Reasonably good agreement was obtained.

Low frequency voltage noise was observed. Hall generator measurements were made to determine $\omega_e \tau_e$.

(Voir page 64 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 71)

THE increased use of electricity over the last decade and the expected increase in the immediate future has made it necessary to look to new methods of increasing the efficiency of converting latent chemical energy, in fuels, into electrical energy. One such method of increasing the conversion efficiency is the magnetohydrodynamic (m.h.d.) method of power generation. Active interest in m.h.d. has been stimulated in the United Kingdom by Thring[1,2] and in the United States of America by Kantrowitz[3,4,5]. From this stimulus the amount of published and unpublished literature on m.h.d. has increased tremendously over the last few years; the greater part of this is concerned with theoretical analyses and not practical experiments. Many excellent reviews[6-12] of this literature are available either by way of individual papers or collected papers at colloquia and symposia. Two of the more important fields for further study are:

(1) The electrical conductivity of the working fluid,

(2) Electrode phenomena such as voltage drop and electron emission.

The work reported in this article is a continuation of studies already made by workers in these two fields on seeded combustion products as working fluid since future practical m.h.d. generators will almost certainly utilize combustion cycles. Hot carbon ablating electrodes and cold water cooled copper non-ablating electrodes were used.

Experimental M.H.D. Generator

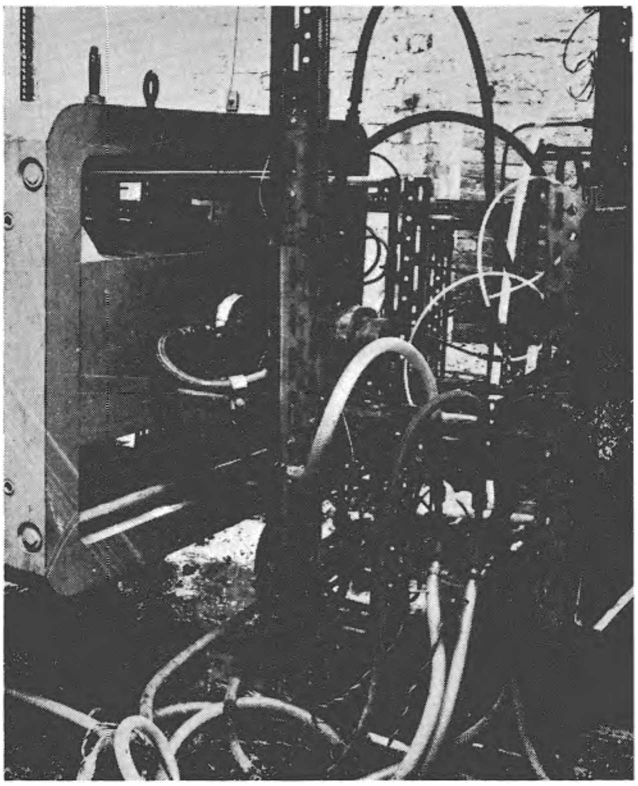
The experimental m.h.d. generator consisted essentially of a water cooled small rocket type meter which operated on gaseous propane and oxygen[13]. See Figs. 1 and 2. The hot combustion gases were seeded with potassium in the form of an aqueous solution of the carbonate and after expansion through a water cooled copper convergent nozzle they were allowed to flow through a ductless magnetic channel of 10kG. This magnetic field was generated by a large permanent magnet of somewhat novel design, Figs. 3 and 4 show drawings of the magnet. Hot carbon or water cooled copper electrodes were situated in the same place as the magnetic field but at right-angles to it. Fc applied electric field experiments the magnetic field wa removed. To assist in the electrical diagnostics the ga temperature was measured with a double beam sodiun line reversal apparatus. See Fig. 5.

Experimental Determination of the Electrical Conductivit of the Hot Stream of Combustion Products

M.H.D. Power Generation

If it is assumed that the hot gas stream has a constar

Fig. 1. Experimental m.h.d. generator



* *Formerly University of Sheffield, now at C.E.G.B.*

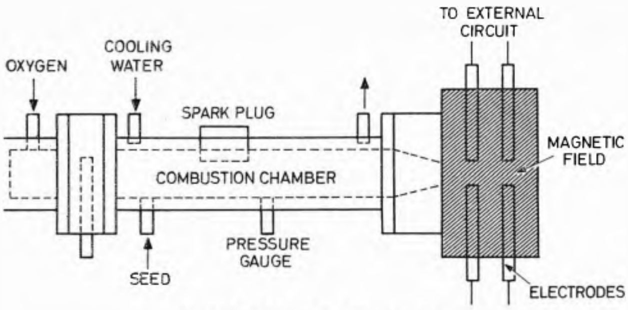


Fig. 2. The m.h.d. generator

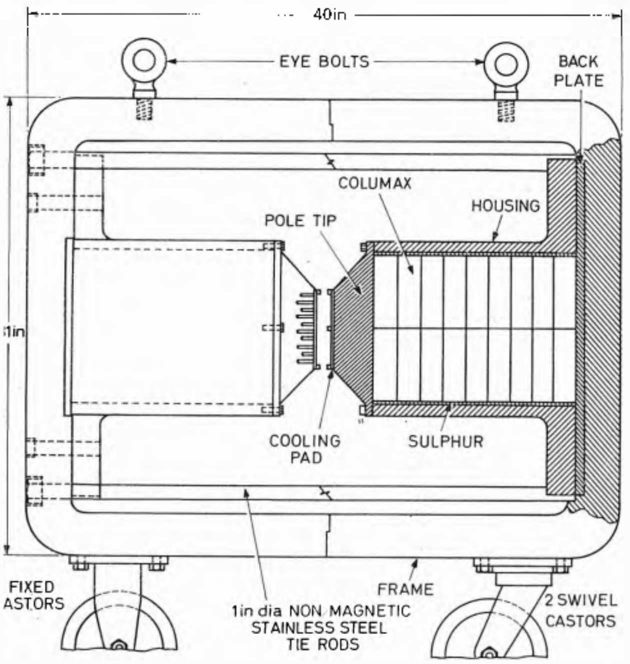


Fig. 3. Permanent magnet assembly. End view

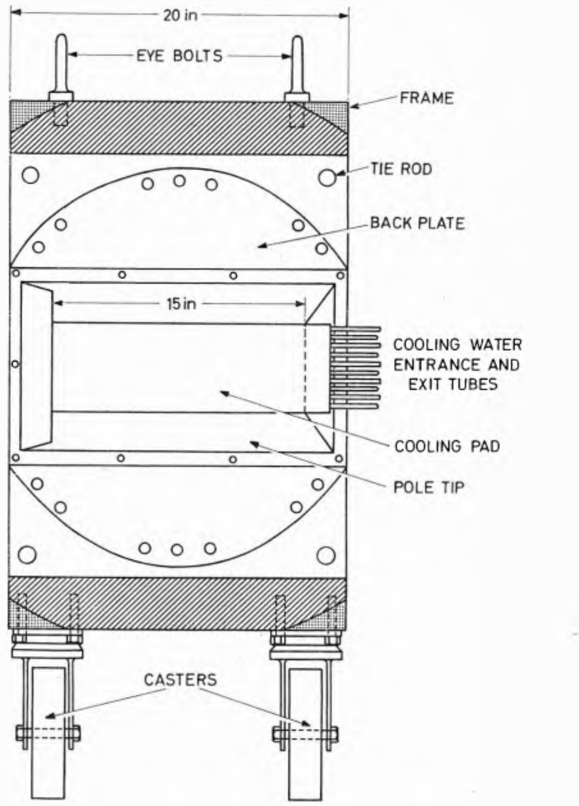


Fig. 4. Permanent magnet assembly. Side view

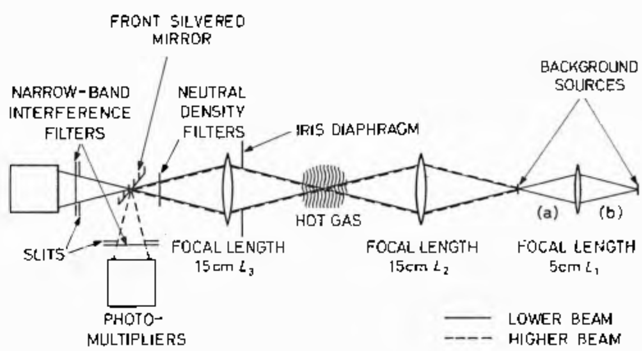


Fig. 5. Optical system for double beam reversal method of temperature measurement

esistance R_g then the relationship between the open-circuit oltage V_{oc}, gas resistance R_g and closed-circuit voltage V_{cc} at current I is given by the equation

$$V_{oc} - V_{cc} = I\,R_g$$

A plot of V_{cc} against I will then give a linear curve f slope $-R_g$ and intercept V_{oc}. From this curve it is ossible to plot the power output $I\,V_{cc}$ against I curve hich will then give a peak when the external load esistance is equal to the generator internal resistance R_g.

Knowing the electrode area A cm^2 and their separation cm it is possible to calculate the specific resistance ρ_0, nd hence the specific conductivity σ_0, of the hot gas tream, from the equation

$$\sigma_0 = 1/\rho_0 = (d/R_g A)\,\Omega^{-1}/\text{cm}$$

This is the conductivity of a column of gas between the lectrodes and is found by using electrolytes to be very lose to the true conductivity, therefore, fringing is negliible (A is large when compared with d).

PPLIED ELECTRIC FIELD

Again, making the assumption of constant R_g then the elationship between R_g and applied V_A and I is given by

$$V_A - V_{ED} = IR_g$$

here V_{ED} is the electrode voltage drop. It is the intercept f a linear plot of V_A against I of slope $+R_g$. The conuctivity can then be determined as already described.

RESULTS AND DISCUSSION

Figs. 6 to 11 show the results obtained from the tests and Table 1 shows the results of experimental conditions. Frost's method[14] was used to theoretically calculate the electrical conductivity collision frequencies and cross-sections.

Good agreement between measured and calculated electronic conductivity was observed for both m.h.d. generation and applied field experiments. The temperature measurements must be treated with some reservation since optical methods measure a mean value only in the zone through which the beams pass and in this series of experiments this was just upstream of the electrodes. Some temperature reduction will occur as the gas stream passes between the electrodes owing to heat transfer and air entrainment. The latter will be small because free jet theory[15,16] predicts that the temperature and velocity will decay only after 3·0 and 3·5 nozzle diameters respectively (2·4in and 2·8in for a nozzle of 0·8in diameter). With both

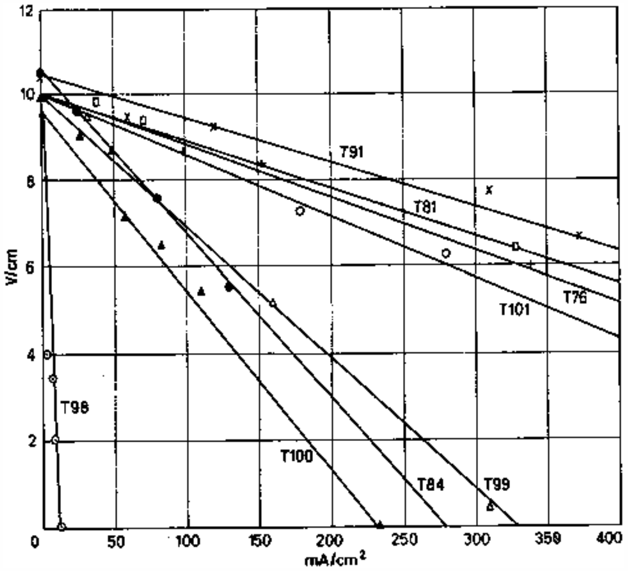


Fig. 6. M.H.D. generator. I-V curves for various K concentrations and temperatures. Carbon electrodes

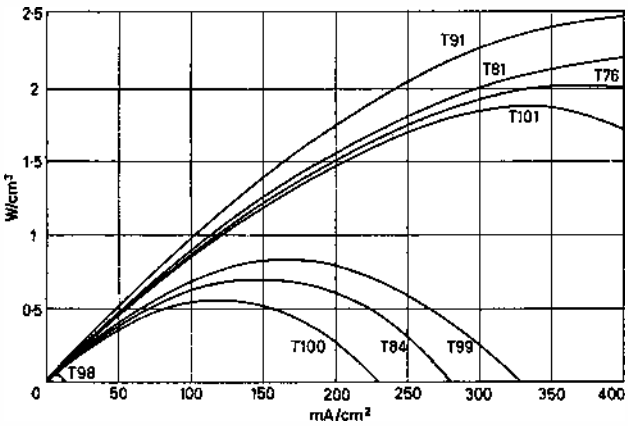


Fig. 7. M.H.D. generator. I-power curves for various K concentrations and temperatures. Carbon electrodes

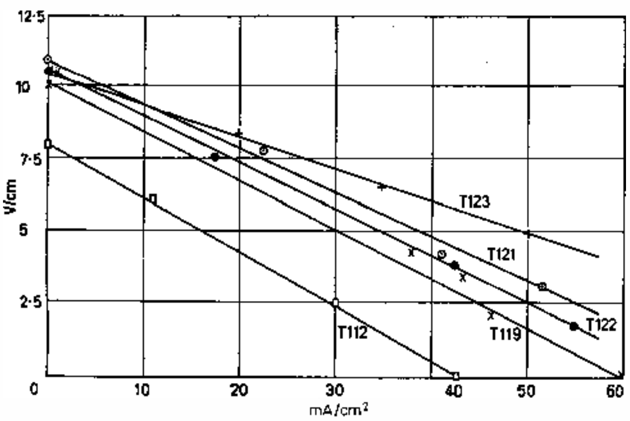


Fig. 8. M.H.D. generator. I-V curves for various K concentrations and temperatures. Water cooled copper electrodes

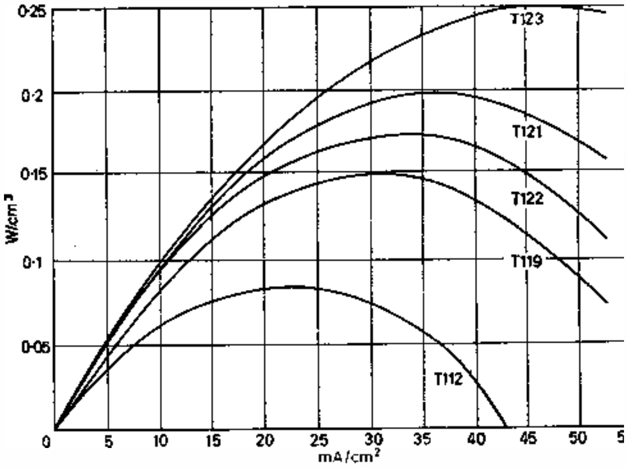


Fig. 9. M.H.D. generator. I-power curves for various K concentrations ar temperatures. Water cooled copper electrodes

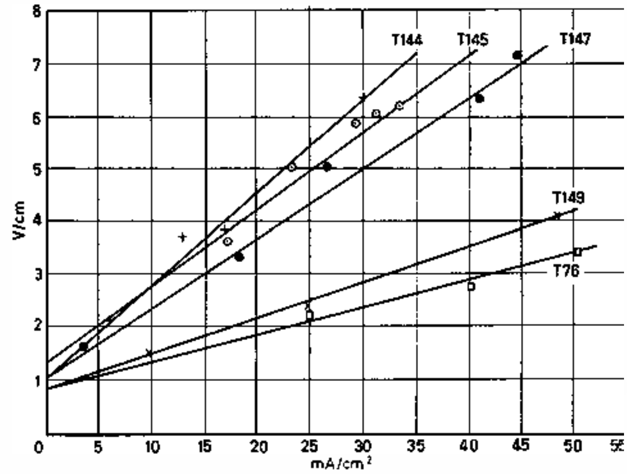


Fig. 10. Applied electric field. I-V curves for various K concentrations ar temperatures. Carbon electrodes

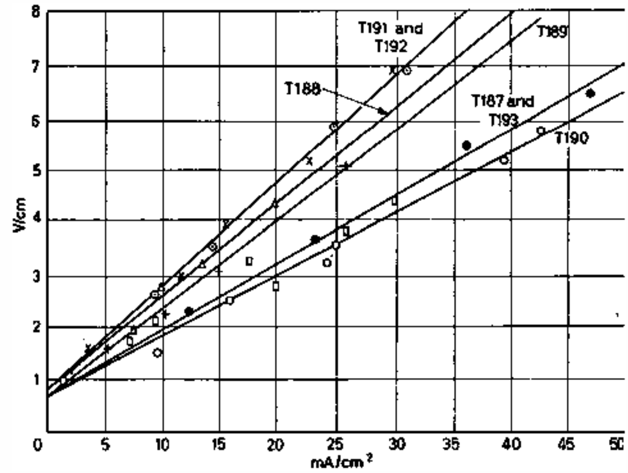


Fig. 11. Applied electric field. I-V curves for various K concentrations ar temperatures. Copper water cooled electrodes

hot and cold electrodes the emitting electrode surfaces should approach the stagnation conditions for the flow and hence boundary layers should not be present; later in the article the surface of the cold electrode will be discussed in more detail.

Agreement between the measured and calculated ele trical conductivity of the gas is very good when it considered that because of its exponential temperatu dependence slight errors in temperature measurement wi greatly affect the calculated values of electrical co uctivity. Electron density and hence electrical conductivity aries very rapidly with temperature

$$\text{Electron density } N_e = K \exp\left(-\frac{3{\cdot}19 \times 10^4}{T}\right)$$

o that at about 1 900°K a 1 per cent error in temperature auses an error of 20 per cent in the electron density.

The measured electrical conductivity (one tenth of the ıeoretical electronic conductivity) for the water-cooled opper electrodes indicates that the electrodes exhibit the nusual phenomenon of electron emission from a cold ırface. This phenomenon has been observed by several ther workers including Swift-Hook[17] and Ricateau[18] with ydrocarbon flames and ieorge and Messerle[19] and ailey[20] in electricity driven ıock tubes in which carbon as not present. In order to xplain this phenomenon it is ecessary to consider the ıteraction mechanism.

When a partially ionized as becomes incident on a ıagnetic field the positive ons drift to one electrode nd the electrons drift to the ther. The high mobility of ıe electrons will make them rift to the anode faster than ıe positive ions will drift to ıe cathode. This separation f charged particles causes a pace charge to be established hich will retard the charged article migration. The properties of the cathode will etermine the mode of electical conductivity of the gas. f it is non-emitting then the ate of removal of electrons t the anode will be limited y that of the ions at the athode and the conductivity vill be ionic. If it is emitting he positively charged cloud lose to the cathode surface vill be neutralized and the onductivity of the gas will e electronic since the elecrons will not be held ack by space charge eparation.

The expected mode of conluctivity of the gas using vater-cooled copper elecrodes would be ionic since he surface temperature of he electrodes will be of the rder of 400 to 600°C and at his temperature thermionic mission will be negligible. An attempt to estimate the onic conductivity of the gas ıas been made and this has een reported in detail elsewhere[13,21]. The absence of nobility data for potassium ons in combustion products makes it impossible to assign a value to the ionic mode of conductivity σ_i but calculation indicates that it must be at least 300 times less than the electronic mode of conductivity σ_e. The experimentally determined conductivity with cold electrodes, 1/10 of the calculated electronic mode, does not appear to be limited by the ionic mode but yet is not as high as that dictated by the electronic mode.

Swift-Hook[17] postulates that the electronic conductivity using cold electrodes originates by electron emission from a carbon layer deposited on the cold metal surface; the carbon of poor thermal conductivity approaching the stagnation conditions of the gas stream. He has slender

TABLE 1

Details of the Experimental Results

RUN NO.	CHAMBER PRESSURE ATMOS. ABS	K (% BY WT)	EQUIVALENCE RATIO	THEORETICAL CHAMBER TEMPERATURE (°R)	(°K)	WEIGHT FLOW RATE (LB/SEC)	(G/SEC)
T.84	1·68	1·10	1·23	5670	3148	0·0788	35·78
T.99	1.65	0·91	1·04	5690	3157	0·0742	33·69
T.91	1·65	0·80	1·01	5670	3154	0·0750	34·05
T.76	1·75	0·885	1·225	5700	3168	0·0740	33·60
T.100	1·65	0·44	1·12	5680	3161	0·0718	32·60
T.98	1·72	Nil	1·19	5700	3167	0·0697	31·64
T.101	1·68	0·695	1·15	5680	3162	0·0723	32·82
T.81	1·68	1·18	1·33	5690	3158	0·0773	35·10
T.147	1·78	1·33	1·52	5600	3110	0·0809	36·73
T.144	1·75	1·23	1·42	5640	3135	0·0787	35·73
T.145	1·75	1·20	1·53	5600	3110	0·0805	36·55
T.149	1·78	1·00	1·28	5700	3167	0·0756	34·32
T.176	1·61	1·14	1·14	5670	3156	0·0743	33·72
T.187	1·65	0·82	0·99	5660	3153	0·0702	31·87
T.188	1·65	0·57	0·99	5660	3153	0·0692	31·42
T.189	1·65	0·70	0·99	5660	3153	0·0699	31·73
T.190	1·65	1.08	0·99	5660	3153	0·0715	32·50
T.191	1·65	1·12	0·99	5660	3153	0·0715	32·50
T.192	1·65	1·33	0·99	5660	3153	0·0728	33·71
T.193	1·65	0·87	0·99	5660	3153	0·0706	32·45
T.112	1·65	1·39	1·00	5690	3159	0·0740	33·60
T.119	1·68	1·66	1·18	5680	3162	0·0772	35·05
T.121	1·75	1·25	1·22	5700	3168	0·0759	33·46
T.122	1·75	0·755	1·22	5700	3168	0·0762	34·60
T.123	1·78	0·543	1·22	5700	3168	0·0745	33·82

TABLE 1 (continued)

RUN NO.	EXIT TEMPERATURE (°R)	(°K)	GAS VELOCITY CALC. FROM CHAMBER TEMPERATURE (FT/SEC)	(CM/SEC × 10⁵)	GAS VELOCITY CALC. FROM O.C.V. (FT/SEC)	(CM/SEC × 10⁵)
T.84	4030	2240*	3 256	0·99	3 500	1·07
T.99	4140	2300*	3 226	0·98	3 350	1·02
T.91	4570	2540*	3 391	1·032	3 480	1·06
T.76	4500	2500*	3 661	1·100	3 350	1·02
T.100	4350	2420*	3 302	1·005	3 210	0·98
T.98	4760	2700*	3 873	1·18	3 350	1·02
T.101	4520	2510*	3 435	1·05	3 350	1·02
T.81	4440	2470*	3 391	1·04	3 350	1·02
T.147	4070	2260	3 493	1·06	—	—
T.144	3960	2200	3 450	1·05	—	—
T.145	4070	2260	3 493	1·07	—	—
T.149	4320	2400	3 579	1·06	—	—
T.176	4450	2470	3 332	1·02	—	—
T.187	4370	2430	3 406	1·04	—	—
T.188	4580	2550	3 493	1·07	—	—
T.189	4600	2560	3 490	1·07	—	—
T.190	4330	2410	3 406	1·04	—	—
T.191	4250	2360	3 362	1·03	—	—
T.192	4250	2360	3 362	1·03	—	—
T.193	4430	2460	3 435	1·05	—	—
T.112	4250	2360*	3 376	1·03	2 630	0·8
T.119	4250	2360*	3 376	1·03	3 340	1·02
T.121	4300	2390*	3 578	1·09	3 600	1·1
T.122	4280	2380*	3 570	1·09	3 510	1·07
T.123	4440	2470*	3 633	1·11	3 500	1·065

* *Estimated gas temperature*

photographic evidence of this carbon layer. Ricateau *et al*[13] has recently produced good photographic evidence of this incandescent carbon layer. Radiation from the body of the gas falling on the cold electrode or turbulence close to the cathode are also possibly explanations of the apparent cold electrode emission[22].

Carbon deposition on water-cooled metal surfaces has been observed in rocket heat transfer studies and attributed to be responsible for the low measured heat transfer rates[23]. By locating a fine thermocouple on the cooled metal surface of the nozzle of a kerosene-oxygen rocket motor Lewis[24] has measured the variation of temperature with time. He found that the temperature fell from 600 to 400°C, suddenly to increase again to 600°C, with a cycle time of 1 to 3sec. Similar tests were made with an amine fuelled rocket motor which would not give carbon deposition. The sawtooth temperature-time graph was not observed proving reasonably conclusively that carbon deposition was occurring in the hydrocarbon fuelled rocket motor. Calculations show that the carbon is approximately 0·010in thick with a hot-side temperature of approximately 1 500°K. The carbon layer is probably amorphous and of high porosity. For a fuel rich film cooled rocket motor Seller[23] suggested that a carbon layer was responsible for the low heat transfer rate. The surface temperature of the carbon was estimated to be between 2 800 to 3 100°K and 0·0022in thick.

Dushman's equation gives the thermionic emission current from a heated surface as follows:

$$Je = A_0 T^2 \exp\left[-b/T\right]$$

where Je is the saturation current density (A/cm²)

T is the absolute temperature (°K)

A_0 is a semi empirical constant (5·93A/cm² deg² for graphite)

$$b = \frac{\phi_1 e 10^7}{k}$$

ϕ_1 is the thermionic work function

e is the electronic charge

k is the Boltzmann's constant

For graphite b is 46 500°K (ϕ_1 is 3·93V). Using the constants a carbon surface at a temperature of 1 500° the thermionic emission is $4{\cdot}23 \times 10^{-7}$A/cm² and for similar surface at 2 260°C the current is $3{\cdot}55 \times 10^{-2}$A/cn Although the emission current is not very high especial at the lower temperature it must be remembered that th temperature is only estimated and also that Dushman equation gives the emission current per unit true or actu surface area. For amorphous highly porous carbon th will, of course, be much larger than the apparent or pr jected area. In the author's experiments there was r evidence of current saturation at values up to 0·4A/cn Also there was no cycling of the conductivity from ion to electronic with a frequency equal to the carbon build-u and flake-off cycle. This is not surprising since if tl mechanism of electron emission is due to the carbc deposited layer it is unlikely that the complete layer w flake-off.

Although the results of the experiments reported in tl article were for fuel rich conditions (i.e. $\phi = 1{\cdot}2$*, corr sponding to maximum flame temperature) the flame did n visibly contain carbon, but carbon deposition could occ by catalytic action of the metal surface or surface oxic The electrodes certainly showed signs of carbon depositic after the test, but this could have ben deposited duri shut-off. According to theoretical equilibrium calculatio the condensed phase carbon should exist until the mixtu ratio is extremely fuel rich, i.e. when the equivalence rat is larger than 3. See Table 2. In practice carbon formati occurs almost immediately the mixture ratio is fuel ri (especially in liquid or solid fuel systems). The mechanis of carbon formation[25] is not fully understood. In t C.E.R.L. experiments the mixture ratio was in the oxyg rich range and again a reasonable explanation would catalystic action.

Electron emission from cold electrodes has been report in the literature. George and Messerle[19] have observ emission from cold electrodes in an air plasma of electrically driven shock-tube in which carbon could n

* *Equivalence ratio,* $\phi = \dfrac{\textit{Mixture ratio, } F/O}{\textit{Stoichiometric ratio, } F/O}$

TABLE 1 (continued)

RUN NO.	MEASURED CONDUCTIVITY (Ω⁻¹/CM)	CALCULATED CONDUCTIVITY (Ω⁻¹/CM)	POTASSIUM PARTIAL PRESSURE (ATMS)	ELECTRON DENSITY (ELECTRON /CC × 10¹³)	COLLISION FREQUENCY MEASURED (SEC⁻¹ × 10¹¹)	COLLISION FREQUENCY CALCULATED (SEC⁻¹ × 10¹¹)	MEASURED COLLISION CROSS-SECTION (CM² × 10⁻¹⁵)
T.84	0·2070	0·029					
T.99	0·0330	0·036					
T.91	0·1000	0·115					
T.76	0·083	0·088					
T.100	0·024	0·052					
T.98	0·001	—					
T.101	0·070	0·080					
T.81	0·091	0·084					
T.147	0·030	0·034	0·0106	4·0	3·76	5·12	3·35
T.144	0·023	0·022	0·0099	2·8	3·47	5·32	3·06
T.145	0·027	0·034	0·0096	3·9	4·13	5·43	3·68
T.149	0·059	0·060	0·0080	6·6	3·16	4·40	2·89
T.176	0·082	0·085	0·0091	10·0	3·43	4·18	3·16
T.187	0·0077	0·066	0·0066	5·4	19·9	4·21	18·4
T.188	0·0055	0·100	0·0046	8·6	44·0	3·63	41·1
T.189	0·0058	0·105	0·0056	9·0	44·0	3·61	41·0
T.190	0·0083	0·068	0.0086	7·0	23·8	4·40	21·7
T.191	0·0049	0·050	0·0090	5·8	33·0	4·66	30·0
T.192	0·0049	0·053	0·0106	6·8	38·7	4·75	35·0
T.193	0·0077	0·075	0·0070	7·6	28·0	4·12	25·9
T.112	0·0054	0·066					
T.119	0·0059	0·071					
T.121	0·0067	0·060					
T.122	0·0062	0·051					
T.123	0·0092	0·065					

ꞔ present. They favour the hypothesis that collision ioniza-on in a thin cathode boundary layer exists when the field rength is high enough (50A/cm^2), a situation analogous › the cathode of a cold-cathode arc.

Bailey[20] has observed the electronic mode of conductivity sing cooled electrodes in a helium plasma produced by ı electrically driven shock tube. Initially the conductivity ionic changing over to electronic after a voltage drop : 9V (the order of voltage drop usually associated with cs).

A variation of the properties of electrodes in m.h.d. enerators has been reported. AVCO have found that a rop of 11V at the cathode and 18V at the anode increas-ıg to eight times these values with varying cooling to the ırbon electrodes occurred[26]. Experimental determination f the electric field in a conducting gas, for both applied ıd induced fields, indicates that the electrode drop is of ual nature: one is associated with charge transfer at the ectrode surface and the second with joule loss as the ırrent passes through the thin aerodynamic boundary ıyer existing over the surface of the electrodes. Experi-ıents show that the electrode drop varies with time, i.e. ith the energy thickness of the boundary layer and the ectrode surface temperature.

Way *et al*[27] with hot carbon electrodes had a voltage rop of only 3V while Swift-Hook *et al*[17] with carbon and ater-cooled copper electrodes and Jones *et al*[28] with ırbon electrodes report *V-I* curves which pass sensibly ırough the origin.

Womack[13] observed small electrode voltage drops—pproximately 1V for the carbon electrodes and 0.5V for ıe water-cooled copper electrodes.

It is not possible to give a satisfactory explanation of ıe apparent electron emission observed from the water-ooled copper electrodes, further elucidation is necessary. is fairly certain that carbon deposition does occur and aving a low thermal conductivity it should approach the agnation conditions of the gas. This together with its igh porosity would give large thermionic emission.

The absence of measurements of voltage drop within the ıs stream together with sheath voltage drop at the elec-odes mean that the results should be treated with some ıution. Good agreement between measured and calculated ectronic conductivity using carbon electrodes was bserved indicating that good electron emission was btained. With the water-cooled copper electrodes the conductivity appears to be electronic rather than ionic, the low value (1/10th of electronic) could possibly be explained by the existence of a boundary layer adjacent to the electrode surface.

TABLE 2

Calculated Equilibrium Condensed Phase Carbon for Propane-Oxygen Combustion for Various Equivalence Ratios at 1 and 2 Atm

CHAMBER PRESSURE 1 ATM			CHAMBER PRESSURE 2 ATM		
EQUIVALENCE RATIO, ϕ	COMBUSTION TEMP. (°K)	CONDENSED PHASE CARBON MOLES	EQUIVALENCE RATIO, ϕ	COMBUSTION TEMP. (°K)	CONDENSED PHASE CARBON MOLES
0.2	2204	0	0.2	2213	0
0.4	2789	0	0.4	2848	0
0.6	2970	0	0.6	3048	0
0.8	3055	0	0.8	3142	0
1.0	3094	0	1.0	3185	0
1.2	3101	0	1.2	3193	0
1.4	3075	0	1.4	3163	0
1.6	3011	0	1.6	3090	0
1.8	2902	0	1.8	2965	0
2.0	2750	0	2.0	2795	0
2.5	2239	0	2.5	2247	0
3.0	1683	0	3.0	1683	0
3.5	1312	0.0213	3.5	1326	0.0214
4.0	1193	0.0815	4.0	1220	0.0824
4.5	1124	0.1385	4.5	1158	0.1396
5.0	1082	0.1916	5.0	1118	0.1922

TABLE 3

Approximate Analytical Expressions for ν_{en}/N for M.H.D. Gases as a Function of Electron Energy in Electron Volts u

GAS	$10^8\ \nu_{en}/N$, sec^{-1} cm^{-3}
H_2O	$10\ u^{-\frac{1}{2}}$
CO_2	$1.7\ u^{-\frac{1}{2}} + 2.1\ u^{\frac{1}{2}}$
CO	$9.1\ u$
O_2	$2.75\ u^{\frac{1}{2}}$
O	$5.5\ u^{\frac{1}{2}}$
H_2	$4.5\ u^{\frac{1}{2}} + 6.2\ u$
H	$42.0\ u^{\frac{1}{2}} - 14\ u$
OH	$8.1\ u^{-\frac{1}{2}}$
K	160

Determination of the Effective Collision Frequencies and the Effective Collision Cross-Sections

For combustion product gases at atmospheric pressure and low seed concentration, the effective collision frequency is determined primarily by the electron-neutral collisions. If the assumption is made that the hot seeded combustion products consist of a single atomic species it is possible to derive the effective collision frequency and hence the effective collision cross-section.

From the measured value of the conductivity σ_0 the electron neutral collision frequency is given by simple kinetic theory as:

$$\upsilon_t = (n_e e^2/m_e\sigma_0)\text{sec}^{-1}$$

where n_e is the electron density calculated by the Saha equation, e is the electronic charge, m_e is the electronic mass.

The effective cross-section for electron-neutral momentum transfer Q_{en} is then given by:

$$\upsilon_t = vNQ_{en}$$

where v is the electron velocity

$$(v = (8kT)/(\pi m_e)^{\frac{1}{2}})$$

k is Boltzmann's constant, N is the number density of atoms.

The experimentally determined values of υ_t were compared with the theoretically calculated values (Frost's method[14]) see Table 3. Table 1 shows the measured and calculated electron-neutral collision frequencies together with the measured electron-neutral collision cross-sections. Close agreement was obtained when it is considered that in the theoretical treatment two major uncertainties exist[28]. The first is that the data used in Frost's theoretical analysis are accurate to only ± 30 per cent. The second is the value assigned to n_e. This electron density was estimated by the Saha equation which only applies to equilibrium conditions. No account has been taken for electron attachment to O_2 O and OH etc; in any case the electron affinities for such electronegative species are not certain. The net effect of negative ion formation is to depress n_e (and increase ion density). An over estimate of n_e by the Saha equation and neglect of negative ion formation would result in an over estimate of the collision cross-section

Q_{en} and collision frequency ν_t. It is also possible for n_e to be increased by a factor such as chemi-ionization, the result would then be to under estimate the values of Q_{en} and ν_t.

The uncertainty of n_e will of course also affect the theoretical value of σ_o. The differences in reported results may be attributed to variations in non-equilibrium factors. Accurate determinations of electron density are required to solve these problems.

Voltage Noise Generation

With a constant magnetic field and gas velocity together with an immovable electrode configuration theory predicts that an m.h.d. generator should generate direct current. Although a steady direct current is indicated by an ammeter there is a considerable noise in the output voltage. Although low[29] (50 to 200c/s) and high[22,30] (1 to 100kc/s) frequency noise has been observed this brief investigation was limited to low frequency noise. The voltage noise generated had a major component of frequency at 104c/s.

In the absence of a detailed experimental programme it is difficult to absolutely explain the noise generation. However, it is tentatively suggested that it could be linked with the low frequency combustion instability known as chugging (50 to 250c/s). This instability appears[31] to be due to combustion delay (of the order of a few milliseconds), which causes adjustment in burning rate consequent on a pressure fluctuation to lag behind it thereby setting up oscillations. These pressure oscillations could then cause the jet diameter to expand and contract, thus periodically immersing the electrodes in and out of the hot gas stream with the consequential generation of voltage noise. This hypothesis is somewhat supported by the fact that the signals from the photomultipliers (spectrum-line reversal apparatus) oscillated with a similar frequency and amplitude, even though the temperature of the gas remained essentially constant with time. Variation in gas stream diameter would affect these signals, but since the two beams were close together they would be affected to the same extent. The absence of suitable instrumentation (such as pressure transducers in the combustion chamber) and the pressure of time made it impossible to thoroughly investigate the phenomenon of voltage noise. It was only possible to make a brief investigation under the normal operating conditions.

Hall Effect Measurements

Simple theory will be used for the determination of the Hall effect[13]:

$$\beta_e = \mu_e B = e\tau_e B/m_e = \omega_e \tau_e$$

the three cartesian components of current density are given by:

$$\left.\begin{aligned} j_{ex} &= (\sigma_o/(1+\beta_e^2))\,[E_{sx} + \beta_e(E_{sz} + uB)] \\ j_{ey} &= \sigma_o E_{sy} \\ j_{ez} &= (\sigma_o/(1+\beta_e^2))\,[(E_{sz} + uB) - \beta_e E_{sx}] \end{aligned}\right\} \;\ldots\ldots\; (1)$$

where j_{ex}, j_{ey}, j_{ez} are the electron current density components, x is the co-ordinate along the duct, y is the co-ordinate in the direction of the magnetic field and z is the co-ordinate in the direction of the electrode pairs. E_{sx}, E_{sy}, E_{sz} are components of the electric field, stationary references frame. ω_e is the electron angular cyclotron frequency (rad/sec), τ_e is the electron-neutral atom collision mean time (sec), B is the magnetic field flux density. Consider the segmented electrodes generator configuration. The open-circuit voltage across any electrode pair is given by the above current density relationships as follows:

TABLE 4

Calculated values of $|\beta_e|$ using Equation (3)

Run No.	100	84	99	101	76	81	91
Short-circuit current (A/cm^2)	0·230	0·230	0·328	0·700	0·836	0·910	1·020
σ_o (Ω^{-1}/cm)	0·052	0·029	0·036	0·080	0·088	0·084	0·115
$\lvert\beta_e\rvert$	1·31	0·60	0·64	0·66	0·61	0·54	0·66

$$J_{ez} = 0$$

$$\therefore\; E_{sz} + uB = \beta_e E_{sx}$$

also

$$j_{ex} = 0 \quad \therefore E_{sx} = -\beta_e(uB + E_{sz}) = \beta_e^2 E_{sx} \;\ldots\; (2)$$

or $E_{sx}(1 + \beta_e^2) = 0$. The open-circuit field is given by uB and the axial field E_{sx} is zero. Secondly consider the case of all the electrode pairs separately short-circuited:

$$J_{ex} = 0 \quad \therefore E_{sx} = -\beta_e(uB + E_{sz})$$

$$\text{but } E_{ez} = 0 \quad \therefore E_{sz} = -\beta_e uB = \beta_e E_{sz}$$

An axial field now exists so there will be a potential difference between the short-circuited electrode pairs. The short-circuit current is given by:

$$j_{es} = \frac{\sigma_o}{1+\beta_e^2}\,[uB + \beta_e^2 uB] = \sigma_o uB$$

without the assumption $E_{sz} = 0$, substituting equation (2) in (1) gives:

$$j_{ez} = \frac{\sigma_o}{1+\beta_e^2}\,[-(E_{sx}/\beta_e) - \beta_e E_{sx}] = -(\sigma_o E_{sx}/\beta_e) \;\ldots\ldots\ldots\ldots\; (3)$$

For the Hall effect measurements the potential difference between two short-circuited pairs of $\frac{1}{2} \times \frac{1}{2}$in surface area carbon electrodes was measured. For an axial separation 0·6cm the potential difference was 3·5V (average of three runs). The conditions for each run were approximately 0·8 per cent K by weight at 2 500°K.

$$\therefore\; E_{sx} = 3{\cdot}5/0{\cdot}6 = 5{\cdot}8\text{V/cm}$$

The open-circuit voltage could not be measured during the same experimental run, therefore an average of E_{sz} of 10·0V/cm, from Fig. 6 was used.

$$\therefore\; \beta_e = E_{sx}/E_{sz} = 5{\cdot}8/10{\cdot}0 = 0{\cdot}58$$

β_e may be estimated in two ways:

(1) By using equation (3) i.e.

$$j_{ez} = -(\sigma_o E_{sx}/\beta_e) \quad \therefore |\beta_e| = (\sigma_o E_{sx}/j_{ez})$$

$|\beta_e|$ has been calculated for several runs as shown in Table 4.

With the exception of Run No. 100, close agreement between the values of $|\beta_e|$ was obtained, with an average value of 0·62.

(2) By using the following approximate equation[26] which applies at 3 000°K and using m.k.s. units for convenience,

$$\omega_e\tau_e = 0{\cdot}06\,B/\rho = \beta_e$$

where ρ is the gas density in kg/m^3 = 0·087kg/m^3 and B is in weber/m^2.

$$\therefore\; \beta_e = 0{\cdot}06/0{\cdot}087 \times 0{\cdot}98 = 0{\cdot}68$$

The measured value of β_e (= $\omega_e\tau_e$) of 0·58 compares very well with that calculated by two different methods, i.e. 0·62 and 0·68. The electronic mobility can now be calculated from the experimental value of β_e

$$\beta_e = \mu_e B = \omega_e\tau_e$$

Again for convenience m.k.s. units will be used

$$\mu_e = \beta_e/B = 0{\cdot}58/0{\cdot}98 = 0{\cdot}59\ \text{m}^2\,\text{V}^{-1}\,\text{sec}^{-1}$$

This value of μ_e agrees reasonably well with that predicted by Frost's theoretical analysis, i.e. 0·4 m^2/V sec.

The electronic mean time between collisions can be calculated using the equation:

$$\tau_e = \beta_e m_e / eB = \frac{0{\cdot}58 \times 9{\cdot}1 \times 10^{-31}}{1{\cdot}6 \times 10^{-19} \times 0{\cdot}98} = 3{\cdot}37 \times 10^{-12}\text{sec}$$

Conclusions

(1) For an m.h.d. generator incorporating hot carbon electrodes the electrical conductivity, effective collision frequencies and collision cross-sections of potassium seeded combustion products were measured at various gas temperatures and seed concentrations in the range 2 200 to 2 500K and 0·4 to 1·6 per cent by weight potassium. The measured conductivities were found to agree well with the theoretically calculated values.

(2) For an m.h.d. generator incorporating water-cooled copper electrodes the gas electrical conductivity was lower than the theoretical electronic conductivity, but nevertheless it appeared to correspond to the electronic and not ionic mode of conduction.

(3) Low frequency voltage noise was observed and was tentatively attributed to low frequency combustion instabilities.

(4) $\omega_e\tau_e$ and μ_e were measured and found to agree well with theoretically obtained values.

Acknowledgment

The author would like to thank Professor M. W. Thring and Drs. I. A. McGrath and R. C. Siddall for their assistance in the above work.

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Radio Emissions from Jupiter

RADIO emissions from the planet Jupiter are still puzzling radio astronomers throughout the world. Attempts to correlate radio emissions to visible features have failed and it is now reasonably established that there is no relationship between the emissions and such visible features on the surface of the planet as the Red Spot, the white spots or any other features.

It was in 1955 that two American radio astronomers, Burke and Franklin, announced detection of signals from Jupiter on 22Mc/s. Signals were sporadic but intense and, unlike other radiations from extra-terrestrial sources, the Jupiter signals were confined mainly to the narrow band of frequencies between 16 and 24Mc/s with a predominance at 18 and 22Mc/s. Signals have, however, been received in Australia at 2·8Mc/s and elsewhere at 38Mc/s and 42Mc/s but these are very rare compared with the number of bursts at 18 and 22Mc/s.

In order to further the investigation—at least seven different theories are current at present—continuous observation over each complete rotation (approximately ten hours) is necessary to avoid loss of valuable data.

A project placed by the National Aeronautical Space Administration with the Florida State University for continuous radio observation of the planet was extended last year to Hyde Radio Astronomy Observatory at St. Osyth, Essex. The principal equipment, shown in Fig. 1, and the aerial system are identical with those installed at Tallahassee, Florida, and the establishment of these two points of observation, separated in longitude by 80°, gives continuous observation over 16-hour periods with a period of overlap when they are observing simultaneously. Burst-for-burst correlations in the overlap period help to eliminate irregularities due to local ionosphere and terrestrial effects.

The major problem in the observations is to distinguish, from continuous pen recordings, the genuine Jupiter signals from those emanating elsewhere. Some twenty types of signal are identifiable, many of them similar to Jupiter.

Until recently the only reliable method of discriminating between the required signal and, in this context, spurious emissions has been by aural monitoring. Aurally, Jupiter signals are characterized by an irregular series of noise bursts constantly varying in intensity and similar to the surging of the sea on the shore. Although, aurally, the signals are readily identifiable by a trained observer there remain the obvious disadvantages of having to have an observer on continuous duty, the possibility of error from fatigue and the number of marginal signals, often of short duration, which leave the operator in some doubt whether to mark a burst as Jupiter or to ignore it. Coupled with this is the natural psychological tendency of an inexperienced operator, during prolonged spells of inactivity of the emissions, to give the benefit of the doubt to a few signals, not appreciating that negative results are scientifically equally rewarding as positive ones.

The necessity for more positive identification has been intensified recently with the establishment of three other out-stations to assist in the observations. These are located in Norway, Nigeria and Spain, roughly on the same meridian as the headquarters out-station at St. Osyth. A further observatory, outside the project and operating independently but supplying information, is located at Rhodes University, Grahamstown, South Africa.

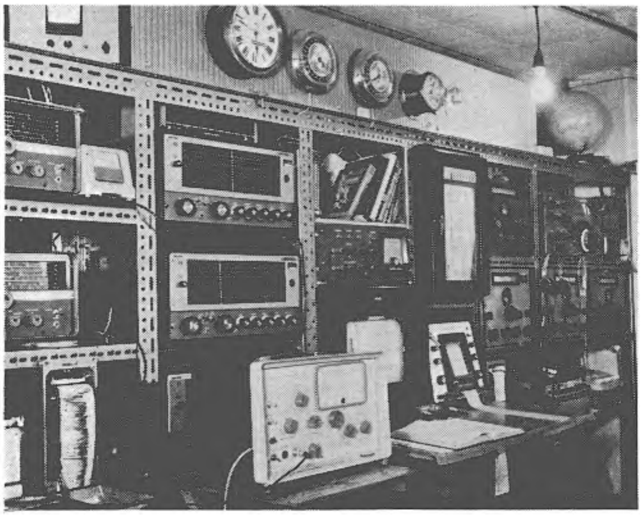
Mr. F. Hyde, owner of the St. Osyth Observatory, believed to be the only full-scale amateur radio astronomy observatory in the world, has developed a simple but effective method of overcoming the difficulties of monitoring.

The method (Fig. 2) uses a hybrid network between the r.f. amplifier in the aerial circuit, and two receiving chains and recorders. Receiver 1 and Recorder 1 are the master receiver and recorder, Receiver 2 and Recorder 2 are a reference receiver and recorder.

In practice the two receivers are initially set up to the required signal frequency by injection of a test signal from a signal generator coupled into the hybrid network. Both receivers are adjusted identically, not only in frequency but also in sensitivity and bandwidth.

The Reference Generator is set to the signal frequency at $100\mu V$ output and very loosely coupled to the aerial

Fig. 1. The receiving equipment at the Hyde Radio Astronomy Observatory
The bays on the left house the receivers and recorders for the Jupiter study. Those on the right are monitoring solar activity on 27Mc/s



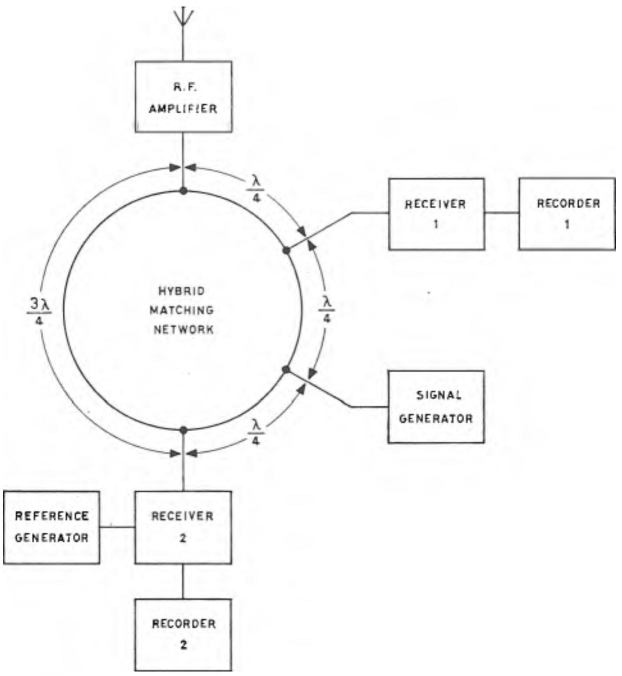


Fig. 2. The hybrid network showing the interconnexion of the major units

circuit of Receiver 2 to beat with incoming signals. Both receivers are used with no a.v.c. maximum usable gain and in the broadest selectivity position. Filters reject all signals below 2kc/s bandwidth. The reference generator coupling is adjusted to provide an injection amplitude equal to the galactic background level, below which no signals are recorded.

The frequency distribution of the signal components in Jupiter bursts is completely random and genuine signals recorded from this source on both recorders are, for all practical purposes, identical. Practically all other signals have a positive or comparatively marked carrier component which beats with the injected signal from the Reference Generator and colours the traces recorded from Receiver 2, leaving the traces from Receiver 1 unaffected.

At the conclusion of an observational period it is necessary only to place the two recordings side by side and make a direct visual comparison to determine genuine Jupiter signals.

The technical requirements of the method are long-term stability, low leakage and extreme reliability in the Reference Generator, in this case an Advance Electronics Ltd SG63D which, although primarily designed as a general purpose f.m./a.m./c.w. signal source has been found to more than meet all demands made upon it while keeping within the limited financial resources available for continued study.

Long-term accuracy of all the receiving, recording and test equipment has also been enhanced by the installation of a heavy duty Advance Volstat constant voltage transformer to eliminate line voltage variations.

The recordings from all stations are sent on a regular basis to Florida State University for computer processing and final results are made available to NASA.

Conclusive results from the Jupiter study are not expected for some time and it may require observations throughout the whole eleven years of a sunspot cycle before clear evidence to support one of the existing theories or, perhaps, a new one, has been obtained.

An Input Transformer with Low Earth Leakage Currents

By D. G. Wyatt*, M.A., D.Phil.

The sources of errors introduced by input transformers used in electrical measuring circuits are investigated and the design of a transformer with low earth leakage currents is given in detail. Particular attention is given to the neutralizing circuits employed. The performance of a complete transformer is given.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

THE usefulness of input transformers in electrical measuring circuits is often limited by imperfections in the transformer itself. Phase angle, leakage reactance, and liability to magnetic interference cause errors not to be found in the "perfect" transformer, while primary/secondary capacitance, asymmetry of primary windings, and electrical leakage from primary to earth limit the degree to which unwanted common mode voltages are isolated from the secondary circuit, as well as causing error in recording the desired voltage.

The effect of leakage impedances Z_1, Z_2, Z_1' Z_2' to earth from the outer ends O, O' and the inner ends I, I' of the transformer primary is observed in general when the source impedance of the applied voltage is finite and when the potential at O, I, O', I' with respect to earth is finite. This does not exclude the possibility of zero potential between the outer connexions O and O'. Thus in Fig. 1 e, e', s, s' are the output voltages and source impedances respectively of a bridge which has one end a of its supply voltage earthed. If $e = e'$ the bridge is truly balanced, but $V_{oo'} \neq 0$ if $S\ (Z_1 + Z_2)/Z_1Z_2 \neq S'\ (Z_1' + Z_2')/Z_1'Z_2'$; that is, a false balance will eventually be obtained. This error can be avoided in principle by using the well-known Wagner earth, with point a not connected to earth, and with O, O' set at earth potential so that no current flows through the leakage impedances.

Fig. 1 is representative also of the input circuit of an electromagnetic flowmeter[1], where $e - e'$ represents the potential difference between points at which the electrodes are placed in the electrolyte and s, s' are the electrode impedances. The values of e, e' may be several hundred times the difference $e - e'$ which is to be measured, and error due to the leakage elements can be very large. Also, since in this particular instance the common mode component in e, e' may be in phase quadrature with the required voltage $e - e'$, and since s, s' have significant resistive components, it is of particular importance that the capacitance elements in Z_1, Z_2, Z_1', Z_2' be greatly reduced.

Reduction of Earth Leakage Currents

The principle which has been used is first to design the transformer so that the leakage reactances Z_1, Z_2, Z_1', Z_2' are as high as possible; and secondly, to make available the connecting points b, b' and to connect these to neutralizing voltages derived from and as nearly as possible equal to the voltages at O, O', respectively. Then Z_1, Z_2, Z_1' Z_2' are ineffective for all values of e, e', $e - e'$ and s, s'.

* *The Nuffield Institute for Medical Research, University of Oxford.*

Transformer Design and Construction

The principle can be applied to any specific design, and that now to be described has been developed as a low-

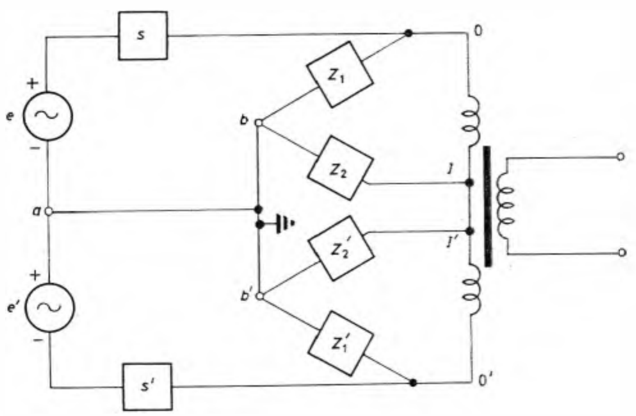


Fig. 1 (above). Representative input circuit

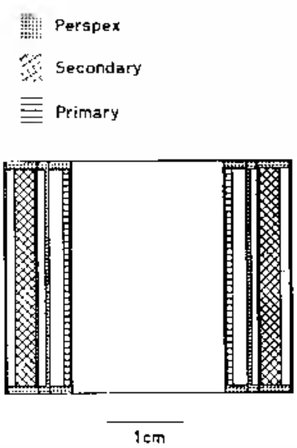


Fig. 2 (left). Axial section of coil formers; the cross-sections are circular

level input transformer for an electromagnetic flowmeter amplifier, operated at a central frequency of 240c/s.

The construction is astatic, with two main bobbins on opposite limbs of a rectangular core composed of 37 layers of L-shaped laminations of 0·015in Permalloy C. (This core material was readily available. A cut core of 0·008in Permalloy C could have been used with advantage.) Each main bobbin consists of four concentric elements machined from Perspex (Fig. 2). The innermost element consists of a bobbin on which is wound one half of the whole primary, consisting of 2·500 turns of 0·0024in Lewmex F wire in seven layers with 0·0015in waxed paper between layers. Fitted over this is the 'intermediate screen', both main surfaces of which are separately coated with silver paint (Johnson Matthey FSP51). The inner surface of this intermediate screen is terminated by a connecting lead at b (Fig. 1), and the outer surface is earthed. The bobbin on which one half of the whole secondary is wound (10 000 turns of 0·0022in Lewmex F wire in 27 layers with 0·0015in waxed paper between layers) is fitted over the intermediate screen. A cylindrical tube is fitted over the secondary bobbin, painted with silver paint, and connected to the outermost surface of the intermediate screen. Thus each half of the secondary is enclosed in an earthed screen, and the secondary terminal pins are screened also. (The painted surfaces are all suitably broken to prevent circulating currents by applying strips of Sellotape 0·5mm wide which are painted over and then removed.) The other main bobbin is similarly constructed, the innermost surface of its intermediate screen being terminated at b' (Fig. 1). Therefore the leakage reactances Z_1 and Z_1' consist mainly of capacitance between the primary outer layers and the inner surfaces of the intermediate screens. These capacitances are kept low by using air-gaps (Fig. 2). Air-gaps are used also between the secondary windings and the surroundings screens, thus reducing secondary shunt capacitance so that the input impedance of the transformer is kept reasonably high. The leakage reactances Z_2 and Z_2' consist of capacitances between the inner layers of the primary windings and the core. The latter is connected either to b or b', so that as at present constructed, Z_2 and Z_2' are paralleled. This simplification is permissible because e and e' are nearly equal. The completed bobbins mounted on the core are shown in Fig. 3, with the primary connectors in the foreground. Impregnation of the complete transformer is to be avoided (especially if neutralization is not used) for this would result in the air-gaps being filled with medium of permitivity ~ 3, with a consequent increase in leakage current. On the secondary side, impregnation would result in higher secondary shunt capacitance, which would of course be unaffected by primary neutralization.

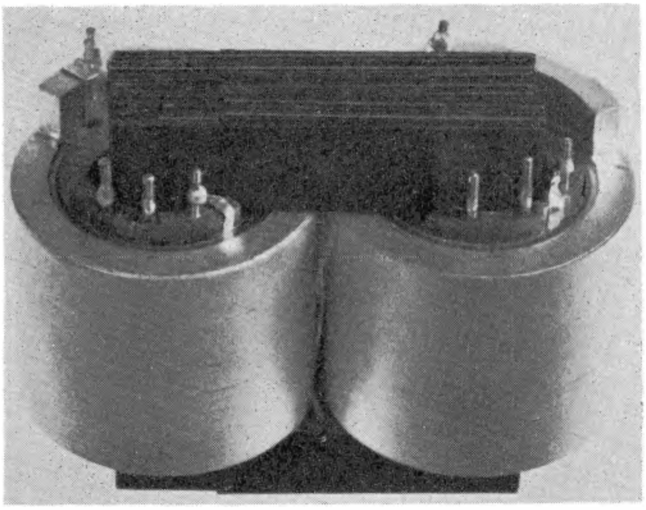


Fig. 3. Completed bobbins and core

However, it is worth noting that it is possible to neutralize the secondary coil/screen capacitances by means of suitable voltages derived from later stages in the amplifier, and thus to obtain a higher input impedance. A transformer designed in this way, using complete primary and secondary neutralization, has been in use for a considerable period.

Transformer Characteristics with all Screens and Core Earthed

The following experimental results refer to the isolated transformer with no connecting leads. The ratio is 1 : 4, primary resistance 2kΩ, secondary resistance 12·1kΩ, primary inductance 300H and primary leakage inductance 0·25H. The stray capacitances, obtained from the resonant frequency and from three-terminal bridge measurements at 1kc/s, are shown in Fig. 4. The capacitance to earth, seen by a common mode voltage looking into either of the primary terminals, is 21·5pF. The capacitance between the surfaces of the intermediate screen is 145pF (C', Fig. 5).

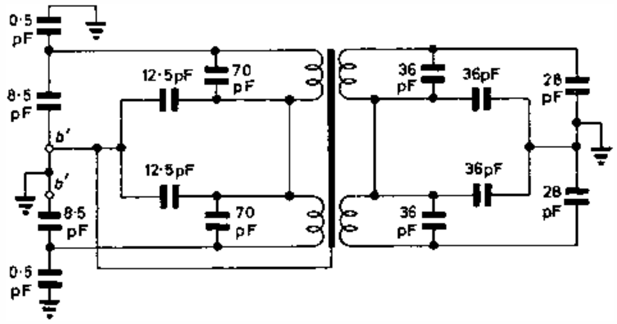


Fig. 4. Effective stray capacitance distribution

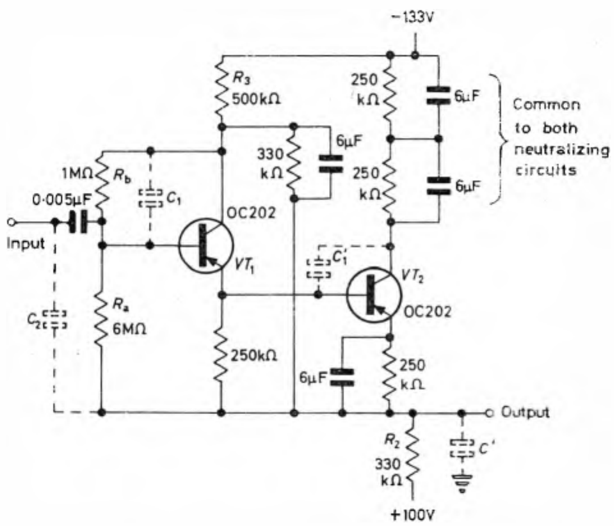


Fig. 5. One of the two neutralizing circuits

The Neutralizing Circuits

The neutralizing circuit is shown in Fig. 5. Each con sists of two emitter-followers[2] connected in cascade. C is the driven capacitance, present between the drive screen and earth, while C_2 is the neutralized capacitanc e.g. that present between the transformer outer leads (or O' and the corresponding driven screen. C_1, C represent depletion layer capacitances. The input capaci tance C_{in} and shunt input resistance R_{in} of each neutrali ing circuit are given by

$$C_{in} = C_1(1 - A) + C_1'/(1 + \beta_1) + C_2(1 - A) + C'/(1 + \beta_1)(1 + \beta_2$$

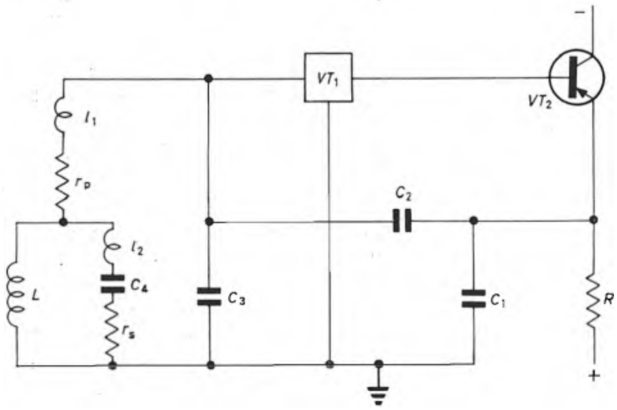


Fig. 6. The oscillatory circuit

$$1/R_{in} =$$

$$\left[(1/R_a) + (1/R_b) + (1/r_{sh1}) + \frac{1}{(1+\beta_1)R_1}\right](1-A) + \left[(1/r_{sh2}) + \frac{1}{(1+\beta_2)\,R_2'}\right]\Big/(1+\beta_1)$$

Where

A = voltage gain from base of VT_1 to emitter of VT_2, β_1 β_2, are the internal current gain factors of VT_1, VT_2, respectively.

r_{sh1}, r_{sh2}. are the corresponding collector shunt resistances.

$R_2' = R_2R_3/(R_2 + R_3)$

Allowing for spread in characteristics of the type OC202, the calculated minimum value of A is 0·9985. Taking $\beta_1 = \beta_2 = 23$, $r_{sh1} = r_{sh2} = 6{\cdot}5\text{M}\Omega$,

$C_{in} = 0{\cdot}0015\ C_1 + 0{\cdot}042\ C_1' + 0{\cdot}0015\ C_2 + 0{\cdot}0017\ C'$ max.

$R_{in} = 59\text{M}\Omega$ min.

Let $C_1 = C_1' = 60$pF, $C_2 = 21$pF, $C' = 145$pF. Then $C_{in} = 0{\cdot}09 + 2{\cdot}52 + 0{\cdot}031 + 0{\cdot}25 = 2{\cdot}9$pF max.

This value of C_{in} is for the transformer and neutralizing circuit alone, without screened connecting leads, and neglects leakage capacitance ($\sim 0{\cdot}5$pF) from the primary connexions to earth. It will be seen that the contribution of the transformer to C_{in} is 10 per cent. Further reduction of C_{in} to a value $\sim 0{\cdot}025$pF could be achieved by using a third emitter-follower in each neutralizing circuit, thus reducing the effect of depletion capacitance. R_{in} would also be increased by a useful factor, to $\sim 300\text{M}\Omega$. The transformer contribution to C_{in} would still be only ~ 20 per cent. When doubly-screened lead is connected to the transformer input and the inner screen neutralized, the contribution of the transformer to C_{in} becomes still less important. As an example, a 6ft length of the doubly-screened lead at present used makes

$C_2 = 300$pF and $C_1 = 400$pF, thus

$C_{in} = 0{\cdot}09 + 2{\cdot}52 + 0{\cdot}45 + 0{\cdot}7 = 3{\cdot}7$pF max.

of which the transformer contribution is 7·5 per cent. If a third emitter-follower were now used, C_{in} could be reduced to 0·9pF, of which 6 per cent would be due to the transformer and 80 per cent to the leads. Thus the indication is that in the present design the primary air-gaps are larger than is necessary when neutralization is used: they could be reduced by a factor of 2 or 3 without seriously increasing C_{in} and the space used to advantage elsewhere, e.g. to reduce secondary shunt capacitance.

Oscillatory Action in the Neutralizing Circuit

When the neutralizing circuits are connected to the transformer, oscillation occurs at a frequency near 20kc/s. The oscillatory circuit (illustrated with a single neutralizing circuit) is shown in Fig. 6. L is the primary inductance, r_s the reflected secondary resistance, C_4 the reflected secondary capacitance, and l_2 the reflected secondary leakage inductance. VT_1 represents the first transistor of the neutralizing circuit as a perfect impedance transformer. The oscillation occurs only under conditions approximating to primary open-circuit; it is prevented, for example, by 100kΩ connected across the primary.

The frequency for which the loop phase shift is zero is given by:

$$\omega_0^2 = (1/2l')\,.\,\frac{C_1 + C_2}{\Sigma C^2}\left\{1 + \sqrt{\left[1 + \frac{4\phi r}{l'^2\,r'\,\Sigma C^2}\Big/\left(\frac{C_1 + C_2}{l'\Sigma C^2}\right)^2\right]}\right\} \quad \ldots\ldots (1)$$

and the circuit is non-oscillatory when

$$\omega_0^2\, l'\, C_3 > 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where $l' = l_1 + l_2 - L/(\omega_0^2 LC_4 - 1) \simeq l_1 + l_2$

$\Sigma C^2 = C_2C_3 + C_1C_2 + C_1C_3$

$r' = (1 - \alpha + (r_e/r_{sh}))\ r_b + r_e$ (referring to transistor VT_2)

$\phi = 1 + (r'/R) + (r_b/r_{sh})$

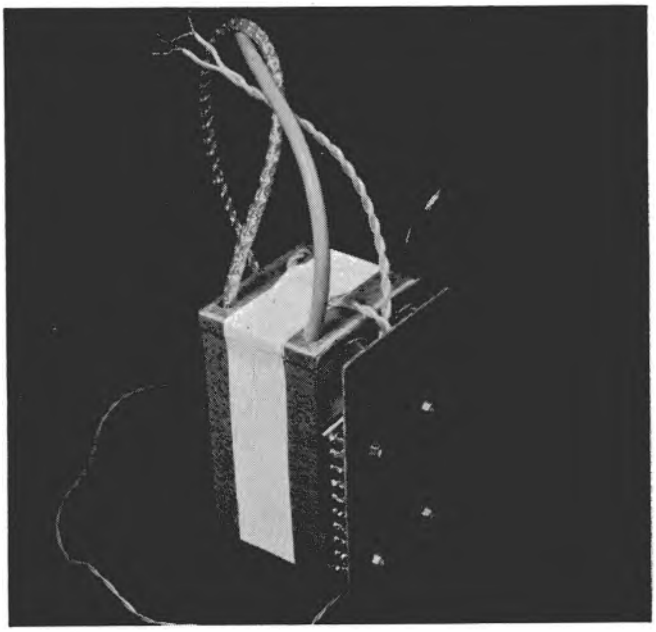
$r = r_p + r_s$

providing that $(r^2/l'^2) \ll (1/l')\,.\,\dfrac{C_1 + C_2}{\Sigma C^2}$, and that

$|C_3r^2 + (C_1 + C_2)\,rr'| \ll |l'(1 - \omega_0^2 l'C_3|$

Prevention of oscillation is basically dependent upon introduction of suitable phase shift by means of C_3, and this possibility is dependent upon the presence of r. If $r = 0$, the non-oscillatory condition (2) cannot be satisfied. Results (1) and (2) give excellent agreement with experiments carried out using a simple inductance l' and values of C_1, C_2, and C_3 such that the oscillatory frequency is not too high, e.g. 6kc/s. However, both results are critically dependent upon phase shift such as may be introduced by the distributed capacitance of the transformer primary.

Fig. 7. Transformer in case, with neutralizing circuits



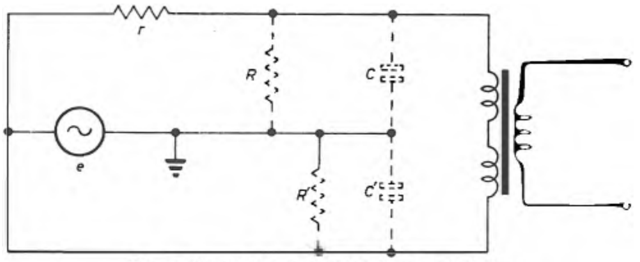


Fig. 8. Circuit for measuring input impedance

Also, when the transformer is connected and the circuit is oscillatory, serious distortion is present which prevents clear definition of the transistor operating conditions, and which demonstrably leads to error in using the formulae. Thus, an oscillatory frequency of 21kc/s is observed, while the value calculated from equation (1) is 33kc/s; and the value of C_3 for which oscillation ceases is 200pF, while the value deduced from equations (1) and (2) is 100pF. A value of 250pF (connected as 500pF across each half of the primary) is used with the present transformer.

Performance of the Complete Transformer

The transformer is enclosed in a Mumetal case, on the outside of which the neutralizing circuits are mounted (Fig. 7). The impedance between each outer terminal of the primary and earth was measured as indicated in Fig. 8. The resistance r is an externally connected resistance, and R, R', C, C' are the apparent shunt resistances and capacitances between each terminal and earth. Application of the voltage e results in a potential difference Δv between the primary terminals, which can be measured at the secondary. Its value is given by:

$$|\Delta v/e|^2 = (r/R)^2 + \omega^2C^2r^2 \quad \ldots\ldots\ldots\ldots \quad (3)$$

When $r/R \ll 1$, $\omega^2C^2r^2 \ll 1$.

By measuring $|\Delta v/e|^2$ at two frequencies, R and C can be found, and R', C' by placing r in series with the other terminal. The measurement is complicated by two factors: first, the voltage Δv which is observed when r is zero, and which is due partly to residual unbalance in the transformer primary and partly to residual leakage capacitance between primary and secondary; second, r is shunted by the input impedance of the transformer. These effects require for their reduction high and low r respectively. However, readings taken with $r = 10\text{k}\Omega$ at 240c/s and 480c/s are calculated to give maximum values of C, C' and minimum values of R, R' with error from these causes not exceeding 5 per cent. When $r = 0$, $|\Delta v/e| = 2{\cdot}1 \times 10^{-6}$ at 240c/s and $3{\cdot}1 \times 10^{-6}$ at 480c/s. The input impedances were found to be:

C_a=2·4pF, R_a=145MΩ; C_b=2·9pF, R_b=150MΩ at 20°C.
C_a=2·2pF, R_a=135MΩ; C_b=3·5pF, R_b=110MΩ at 45°C.

Increasing C_2 and C' (Fig. 5) by 330pF and 470pF respectively in both circuits increases C_a to 3·2pF and C_b to 4·3pF. This is for the complete transformer with 1ft of screened and neutralized connecting leads on both outer and inner terminals of the primary, and with the unit enclosed in an outer Mumetal case.

The input impedance was measured by connecting in series with each side of the primary a capacitor of value 2200pF, and applying between the open ends a voltage of 300μV r.m.s. balanced with respect to earth. The phase and quadrature components of voltage across the transformer primary were measured by means of an accurate phase discriminator. The effective input impedance at 240c/s consisted of 0·9MΩ shunted by 100pF. The resistive component of impedance is due almost entirely to eddy current loss, and agrees well with the calculated value.

Extension of the Method

It is possible to neutralize the capacitance between the secondary windings and the surrounding screens by feeding the latter from voltages as nearly as possible equal to the mean potential of the adjacent layer of winding. This results in lower capacitance reflected into the primary and higher resonant frequency. It may result also in oscillation similar to that which occurs with the neutralizing circuits, but which may be prevented by connexion of a suitable capacitance across the secondary terminals. This limits the advantage to be gained, but is not always necessary. The analysis is similar to that of Fig. 6, and in the one instance examined (a transformer with leakage capacitances considerably greater than those of the transformer described here) gave results very close to the experimental values, both for oscillatory frequency and for critical shunt capacitance.

The possibility exists also of improving performance by using fed screens which are distributed in a winding, thus reducing the effective winding capacitance.

Acknowledgments

The author is indebted to Messrs. Parmeko Ltd for their co-operation in the initial stage of this development, and to Dr. R. M. Glaister, of Standard Telecommunication Laboratories Ltd for data concerning low-level losses in Permalloy C. The transformer was expertly manufactured by Mr. H. C. Holt.

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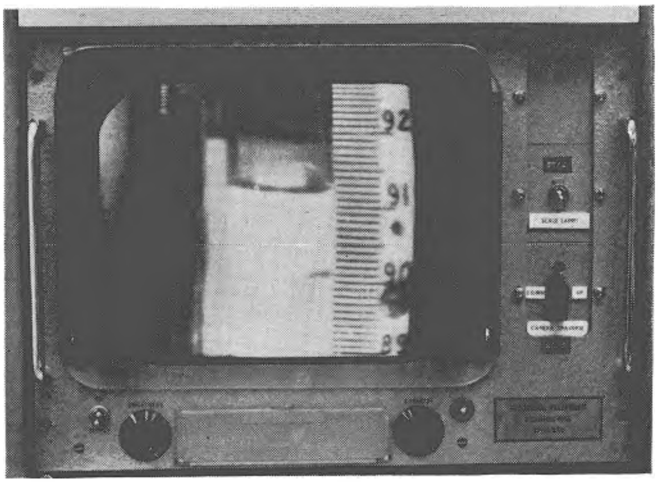
Reactor Inspection by Television

Information regarding the liquid level inside Helen III nuclear assembly at the Atomic Energy Establishment, Winfrith, Dorset, is transmitted to a television receiver outside the reactor cave housing the facility, within an accuracy of ±0·5mm, by a closed-circuit television system supplied by EMI Electronics Ltd.

Closed-circuit television is one of two independent systems used for transmitting information on the level of the liquid moderator from inside the reactor cave. Variations in the height of the moderator are magnified by the camera and so more easily seen on the television screen.

An EMI type 6 minicamera is mounted at right angles to the gauge and is driven by an electric motor so that the camera rises and falls in sympathy with the liquid moderator and is therefore always viewing the liquid level in the gauge. Details of this level are immediately transmitted to the monitor outside the Nestor cave.

Liquid moderator level shown on television monitor



A Neurophysiological Computer

By B. Delisle Burns*, W. Ferch* and G. Mandl*

Due to an increasing interest in the time-average behaviour of nerve cells, there is a growing demand for instruments that will rapidly provide an on-line classification of time intervals. A relatively inexpensive, special-purpose computer is described, which will perform interval analyses, and digital, cross- or autocorrelation.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

It is not hard to justify the development of a relatively inexpensive, on-line, averaging computer for neurophysiological research. During recent years, there has been a steadily increasing interest in the time-average behaviour of, either groups of neurones, or single nerve cells[1-6]. The present authors have been mainly interested in recording the times of discharge of single neurones with extra-cellular micropipettes, and have previously designed two other special purpose computers for analysing such records[7,8]. Both of these instruments suffered from certain disadvantages. For this reason the design was undertaken of the computer described below. The objective was to construct a relatively cheap instrument that could easily be programmed in a variety of ways and that could perform on-line analysis during a biological experiment. Naturally, the problems of cost and required accuracy had to be considered together; there is no sense in obtaining a measure which is accurate to ± 1 part in 10^6, if most users would be happy with ± 1 per cent. Therefore an attempt was made to use the cheapest reliable design that would just provide an order of accuracy that seemed acceptable for the common forms of biological analysis.

The resulting instrument can be programmed in four different ways, all of which involve the classification and counting of time intervals. Two identical input channels (1 and 2) feed the computer through voltage gates with digital signals, while the computed output displays the counts accumulated in 100 equal storage bins, each bin of a chosen duration δT. The four modes of operation are illustrated in Fig. 1 and can be listed as follows:

(1) *Digital-autocorrelation* (Fig. 1(a)) = the frequency distribution or total numbers of signals on channel (2), that follow *any* preceding signal after times lying between T and $T + \delta T$, and which are not greater than $100 \times \delta T$. (δT must be smaller than the least interval between signals). This form of analysis is appropriate for the detection of tendencies toward cyclic behaviour. It would answer such questions as: 'Does this neurone occasionally fire in bursts occurring at a fixed frequency?'; 'Does the heart exhibit occasional auriculo-ventricular block?'.

(2) *Cross-correlation* (Fig. 1(b)) = the frequency distribution of signals on channel (2) that follow *any* preceding signal on channel (1) after times lying between T and $T + \delta T$, and which are not greater than $100 \times \delta T$ (δT must be smaller than the least interval between signals on channel (2)). Cross-correlation will indicate for example, the probability with which a discharge of one neurone follows the discharge of a neighbouring nerve-cell; alternatively, this form of analysis could indicate the extent to which timing of the cardiac beat is modified by respiratory inspiration.

(3) *Interval-analysis* (Fig. 1(c)) = the frequency distribution of intervals between *consecutive* signals on channel (2), lying between T and $T + \delta T$, and which are not greater than $100 \times \delta T$. Interval-analysis can be used to identify the most common interval between consecutive discharges of a neurone, or to

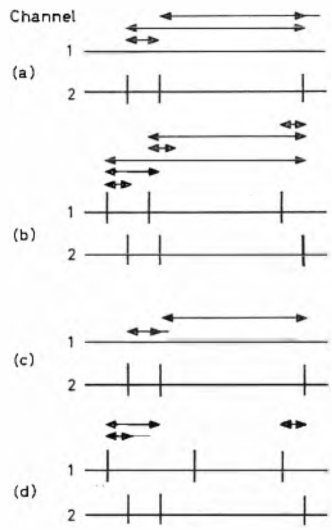


Fig. 1. Illustrating the time intervals (⟷) that are measured and classified by the computer, in its four available programmes

(a) Digital autocorrelation; (b) digital cross-correlation; (c) interval analysis; (d) post-stimulus histogram

detect the functional refractory period of that neurone. The same form of analysis could also be used to detect the functional refractory period of any *AV*-bundle in a case of auricular fibrillation.

(4) *Post-stimulus-histogram* (Fig. 1(d)) = cross-correlation as in (2) above, where the intervals between signals on channel (1) are all equal (cyclic). The computed output may be set so that it is distributed between any chosen number of bins from 1 to 100. This programme is useful for determining the relation between a cyclical stimulus and the average biological response.

The computing section of the instrument handles digital information, while its memory is analogue and consists of potentials stored within a bank of 100 capacitors. Fig. 2 shows a simplified block diagram of the instrument, complete with its continuously rotating commutator (read-out) which can feed results to either an oscilloscope or a pen-writer. It will be seen that the measurements of time referred to in Fig. 1 are classified by means of a 100-step shift-register. The shift-register operates a series of 100 gates, which permit charging pulses to pass into the appropriate capacitors only when the gates are open (when the equivalent switches shown in the diagram are closed). The transfer speed of the shift-register is controlled by an adjustable clock-circuit, providing pulses at a frequency

* McGill University, Montreal, Canada.

of $1/\delta T$. In Fig. 2 the clock includes both clock and binary count-down of Fig. 4. The property of such a shift-register is that if the n^{th} gate is open, a pulse received from the clock will shut that gate and open the $(n+1)^{th}$ gate. The first gate of the shift-register is opened by a signal, chosen from the two available inputs by the operator; the next pulse from the clock closes this gate and opens the second, which remains open for a time δT until the next clock-pulse again shifts the position of the open gate one further down the series of 100 gates. Thus the shift register carries an analogue (in terms of open gates) of the timing of signals provided to the first gate, in their proper temporal sequence, for a time $100 \times \delta T$.

The circuit diagram and manner of operation of one of the 100 memory units is shown in Fig. 3. The count for each bin is accumulated and stored as a potential across the capacitor C. At the top of the figure is shown an input to channel (2), representing spike discharges from two cells recorded from an extra-cellular microelectrode. The voltage gate for this channel is presumed to have been set so that only the larger signals are counted by the instrument. The output from the voltage gate (Fig. 2) generates charging pulses of fixed magnitude and a duration which can be preset by the operator. It is the duration of these charging pulses that determines the full scale count of the instrument. These charging pulses are supplied in parallel through individual resistors R_1 to all gates, but for any gate that is shut they are shunted to earth through the diode MR_1. Should they arrive at an open gate, the charging current passes through R_2 and the diode MR_2 into the capacitor C, causing a step-increase in stored potential. The diode MR_2 must provide a very small reverse current since the useful storage-time of the instrument will depend upon the quality of this diode. With $C = 1\mu F$. 1 per cent of the stored potential will be lost in 10min if the reverse resistance of MR_2 is $10^{11}\Omega$. (An IN3575 has been used). During operation, the capacitor C charges asymptotically to a potential of $-12V$. When a rectilinear representation of the counts in each bin is required, full scale is limited to that count providing a charge of 1V across C; it can be shown that in these circumstances the maximum non-linearity of the recorded counts is less than 1 per cent of full scale. The resistor R_2 is necessary to ensure that all bins (capacitors, C) accumulate the same potential for the same count; it would be impossible to ensure this, without adjustment of R_2, unless the forward resistances of the 100 diodes, MR_2 and the capacitances of the 100 capacitors C were exactly equal. (The proper setting of the 100 values of R_2 is not as tedious as it might appear; a nul-method is employed, and the 100 bins can be set to equal sensitivity $\pm 0{\cdot}2$ per cent in about half an hour. Once set, they should not need readjustment for more than a year of daily operation.)

In order to illustrate the operation of the instrument as a whole, consider the calculation of a post-stimulus-histogram, such as that shown in Fig. 5(b). The original record[2] of Fig. 5(a) is a short sample from a three-minute recording obtained from a nerve-cell in the visual receiving area of the cat's cerebral cortex, while the animal was 'observing' a simple pattern of light and dark areas, focused upon its retina and made to oscillate up and down at 3c/s with an amplitude of 0·5°. Programming the computer for post-stimulus-histogram will provide an answer to the question: 'On the average, how often did this cell fire at various times after the downward movement of the pattern ?' The record of neuronal activity is

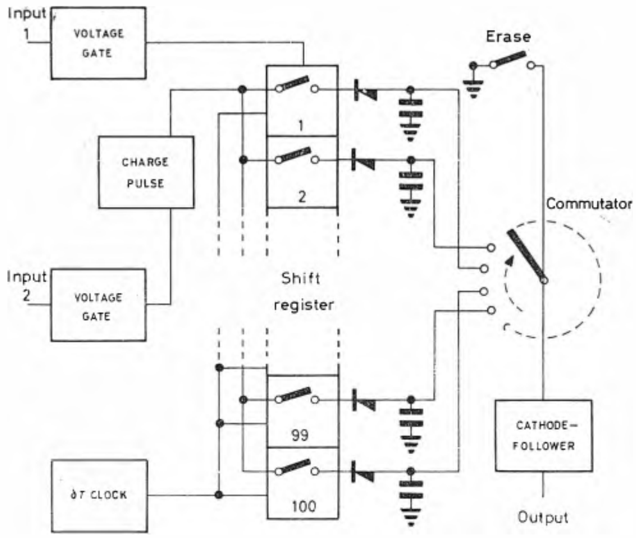
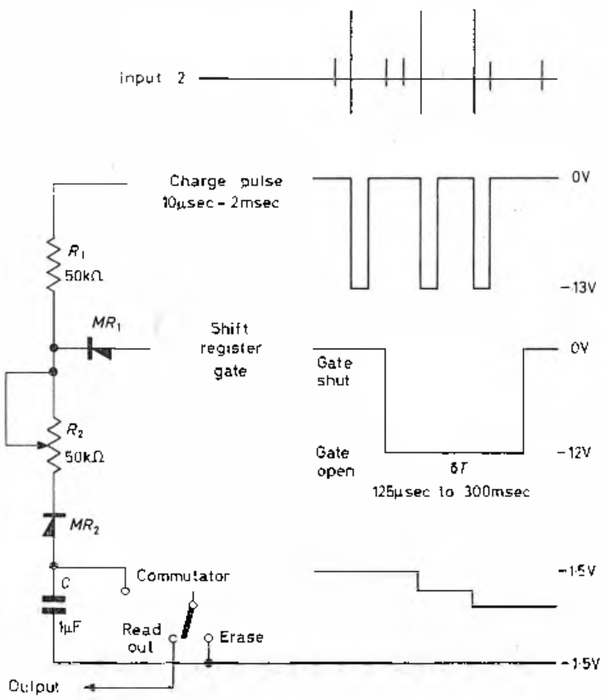


Fig. 2. A simplified diagram of the computer

supplied to channel (2) and the voltage gate is adjusted so that each action potential is counted and no unwanted back-ground signals are scored. The record of the times of pattern movement is supplied to channel (1) and the voltage gate is set so that only those signals indicating downward movement are admitted. Thus, each downward movement of the pattern opens the first gate of the shift-register; at the same time the clock is 'reset to zero' so that the first gate stays open for a time δT. If a record of the behaviour of the neurone is required, covering the whole period between identical stimuli, then the clock-speed would be adjusted so that $100 \times \delta T = 1/F$, where F is the frequency of stimulation (as in Fig. 5(b)).

Sometimes, when the number of signals on the second channel is small, it is convenient to accumulate the counts in fewer than 100 bins. Say 30 bins were regarded as

Fig. 3. Illustrating the memory of the instrument

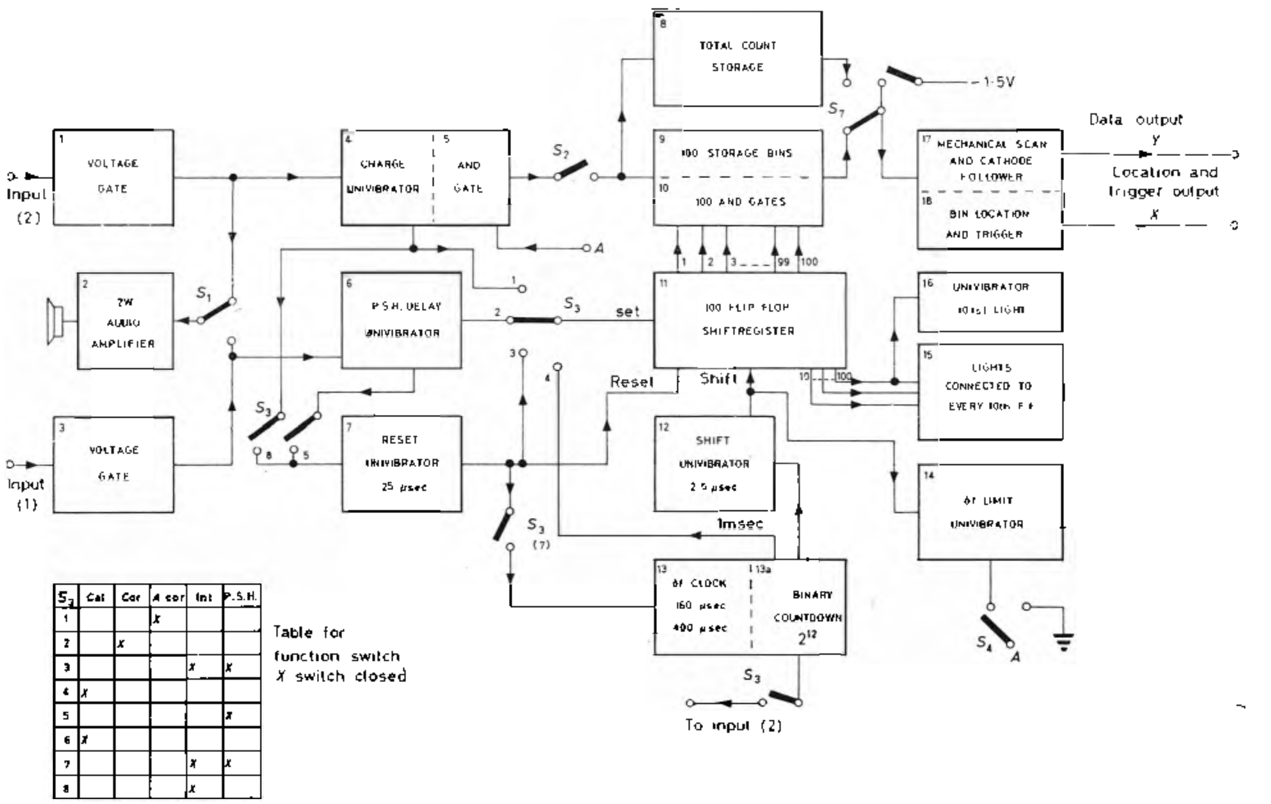


adequate, then the clock-speed would be set so that $30 \times \delta T = 1/F$. This is made possible by the fact that each signal on channel (1) not only resets the clock and opens the first gate, but first 'clears the shift-register' (i.e. closes any gates that may still be open). If the response to be examined is all over within 100msec of the pattern movement, then the clock-speed might be set so that $\delta T =$ 1msec. (100 $\delta T =$ 100msec). This provides for maximum temporal resolution of the response.

Suppose that δT has been set to 1msec, then every signal that comes through on channel (2) during the third millisecond after a downward pattern-movement, will be

to preset R_2 for the total-count circuit, so that its count sensitivity is 1/50 that of the 100 gated bins.

In order to explain in detail the four modes of operation of this computer, a functional block-diagram of the instrument is provided in Fig. 4. A complete wiring diagram has not been given, since no novel circuits are used and many circuits exist, that would provide the necessary functions. The purpose here is only to provide a description of the logical construction that has been found essential. In Table 1 are given the connexions between the elements of Fig. 4 that are necessary for the various modes of operation.



S_3	Cat	Cor	A cor	Int	P.S.H.
1			X		
2		X			
3				X	X
4	X				
5					X
6	X				
7				X	X
8				X	

Fig. 4. The full diagram of the computer

counted as an increment of charge on the third capacitor, C of Fig. 3. In general, any action-potential occurring between $(n-1)\,\delta T$ and $n\,\delta T$ will be stored in the n^{th} bin. In this way, after a period of analysis, the output of the instrument will indicate the temporal distribution of action-potentials following the average downward pattern-movement. The required sensitivity of the ordinate in the display is obtained by adjusting the duration of the charge-pulse (Fig. 3). When using linear display, full-scale (=1V) can be preset to values from 5 counts to 1 000 counts per bin. If the storage capacitors of Fig. 3 are allowed to charge beyond 1V, an exponential scale is available providing half-full-scale = 6V representing a maximum of some 8 000 counts; 12V represent an infinite count.

The total count of signals on channel (2) is accumulated in a similar way to that illustrated in Fig. 3, on a capacitor $C = 20\mu F$, which is supplied by an ungated charging pulse; thus, every charge-pulse gets through to this capacitor. The charge on the capacitor may be read through an intermediate position (not shown in Figs. 2 and 3) of the erase switch which connects the read-out cathode-follower manually to this capacitor. It has been found convenient

In Table 2 are given the specifications of the various functions illustrated in Fig. 4.

In order to use the computer properly, one must know its limitations. Reference has already been made to the fact that ordinates of the output represent counts with a maximum error of ± 1 per cent of the full-scale count. Moreover, values of δT and full-scale counts cannot be preset to better than ± 1 per cent of a chosen value. These errors are small by comparison with the total variability intrinsic to the majority of biological phenomena that the computer is designed to analyse. More important to the user, is consideration of the errors in computation imposed by the instrument's logical design and the limitations of its operation. For example, the charge-pulse has a finite duration; thus, a charge-pulse that is 1msec in duration will fall into 8 bins, when δT is 128μsec. Therefore, in all modes of operation, the charge-pulse must be kept shorter than the chosen value of δT, so as to minimize temporal misplacing of counts. In any case, complementary fractions of one count can always enter neighbouring bins. Some of the other intrinsic limitations and errors are indicated in Table 3.

The description given above has referred exclusively to the analysis of digital forms of biological information—the times of 'spike discharge' recorded from single neurones with an extra-cellular microelectrode. In Fig. 5 are illustrated various results computed from a record of this type. Since the input voltage gates are direct-coupled, the instrument can be used for analysis of some continuously variable biological waveforms. Thus, by supplying input (1) with an electrical measure of chest movement, while input (2) is supplied with the electrocardiogram, one could perform a cross-correlation between respiratory movement and heart rate. Likewise, by supplying the electroencephalogram to channel (2) and the times of sensory stimulation to channel (1), the probability distribution of e.e.g. voltages in excess of a chosen value following the average stimulus, could be determined.

TABLE 1

The Operations Performed Sequentially by the Two Input Channels of Figs. 2 and 4.

R = clear shift register; i.e. close any open gates, reset clock and binary count-down to zero

C = initiate charge-pulse

O = open first gate of shift-register

	DIGITAL AUTO-CORRELOGRAM	DIGITAL CROSS-CORRELATION	INTERVAL ANALYSIS	POST-STIMULUS HISTOGRAM
Channel 1	NOT USED	*O*	NOT USED	*R.O.*
Channel 2	*C.O.*	*C*	*C.R.O.*	*C*

Note: When only channel (2) is in use, the charging pulse is complete before the next operation begins.

TABLE 2

FUNCTION	SPECIFICATIONS
Voltage gates	Fed direct coupled from inputs (1) and 2, providing input impedance of 50kΩ. Count all signals more positive than any chosen value between −6V and +6V. Hysteresis = 0·2V.
Speaker	May be connected to the output of either voltage gate, providing clicks of adjustable volume whenever chosen gate responds to its input.
Charge univibrator	Provides chosen durations of charge-pulse (Fig. 2) from 10μsec to 2msec, giving linear full-scale counts from 5 to 1 000.
Clock	Provides cyclic signals at a chosen, continuously variable, frequency between 16kc/s and 6kc/s.
Binary count-down	Provides binary count-down of clock-frequency for the determination of δT. The number of binary steps used may be chosen so that δT may have any value from 128μsec to 300msec, and may be changed in accurate binary steps.
Calibration system	Holds all gates continuously open and provides charging pulses at frequencies selected from clock and count-down combination.
Delay Channel 1	Available during post-stimulus histogram operation. Provides a delay between signals on channel (1) and operations *R* and *O* of Table 1. Values of delay between 0 to 500msec.
δT limit, *L*	Available during auto- and cross-correlation. Arranges that gates open only for the first fraction (*L*) of δT; this permits the use of δT values that are longer than the shortest intervals between successive signals fed to the first gate.

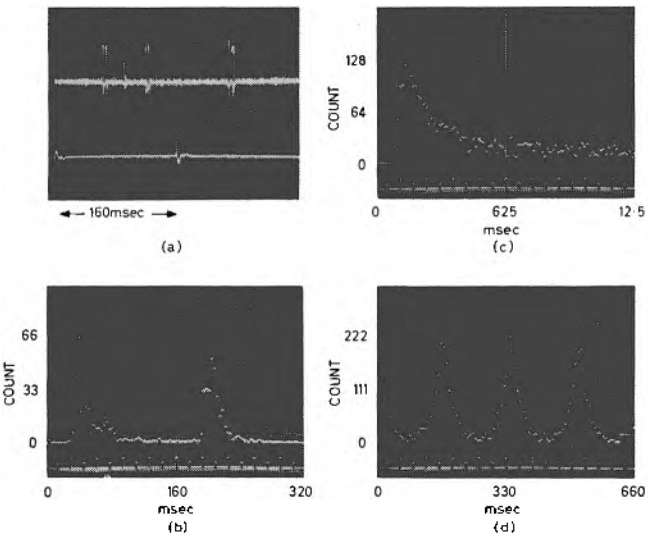


Fig. 5. A sample of original, (retouched) two channel record

(a) Described in text, and three different analyses of this record; (b) = post-stimulus histogram; (c) = interval analysis; (d) = digital autocorrelogram

Much greater flexibility in the analysis of continuously variable waveforms can be obtained by the use of a frequency modulator, which converts the continuously variable input into digital signals with a frequency that is proportional to the input voltage. These digital signals can then be fed to the computer in the ordinary way. The best form of frequency-modulator for use with the computer described above is one providing a relatively large modulation of the mean carrier frequency. It has been found convenient to use a modulator that is linear (within 1 per cent) over the range (−2V = 200c/s) to (+2V = 2 000c/s). Thus, using a mean carrier of 1kc/s, and allowing peak modulations of 200c/s and 1 800c/s, one can average responses to stimulation at 1/sec (a complete post-stimulus histogram) for a period of about 2min. If the interval containing the response is only 1/5 of the 1/sec interval between stimuli, then records of some 10min duration can be averaged.

Conclusion

As already stated the purpose was to design a relatively

TABLE 3

Limits of Operation of the Computer

C = charge-pulse duration = (5/full-scale count) msec

L = δT-limit

I = least interval between signals fed to the first gate

PROGRAMME	C	δT	MAXIMUM FREQUENCY ACCEPTABLE Ch (1)	Ch (2)
Digital auto-correlogram	$C<\delta T$	$\delta T<I$ If $\delta T>I$, set $L<I$		$1/2C$
Digital cross-correlation	$C<\delta T$	$\delta T<I$ If $\delta T>I$, set $L<I$	$1/2\delta T$	$1/2C$
Interval analysis	$C<\delta T$			$1/2C$
Post-stimulus histogram	$C<\delta T$			$1/2C$

Note: In all cases, the most probable placing of counts will be C/2 late; the maximum temporal displacement of recorded counts will be +δT, or 1 bin, late.

cheap, special purpose, averaging computer for biological data, that could be used on line. Accuracy has necessarily been traded for economy; its limitations have been dealt with at some length. Despite these limitations two years employment of this computer in the solution of a variety of biological data has convinced the authors of its usefulness and reliability.

Acknowledgments

The development of this instrument was supported by the Canadian Defence Research Board and the National Research Council of Canada, to both of which the authors express their thanks. They would also like to thank Professor J. G. Robson who gave much valuable advice.

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Zero Phase-Shift Multivibrators

By A. Pugh*, B.Sc.

Investigations into circuits for handling positive and negative going pulses has led to the construction of transistor multivibrator circuits using grounded-base stages, as opposed to grounded-emitter stages. The article draws attention to the rather novel properties of a series of circuits designed on the above principle and indicates some applications.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

THREE fundamental requirements need to be satisfied for a conventional (two stage) multivibrator to operate, viz:

(a) The loop gain must be greater than unity.

(b) The total phase shift around the loop must be zero.

(c) One stage must be driven so that it becomes conducting, while the other stage, at the same instant, is driven so that it becomes non-conducting.

It so happens that in conventional multivibrator circuits, all three requirements are automatically satisfied by cascading two similar stages of amplification, each having a phase shift of 180°. If attempts are made to meet requirements (a) and (b) by cascading two similar stages of amplification, each now having zero phase shift, both stages would be conducting or non-conducting simultaneously, and requirement (c) is then no longer satisfied and the circuit will not operate.

However, if one of these stages consists of a grounded-base pnp transistor and the other of a grounded-base npn transistor, then requirement (c) can be fulfilled. Fig. 1 shows the basic circuit of a bistable multivibrator designed in the light of the above reasoning; however, this circuit is not practicable because it is not possible to derive from each collector circuit sufficient current to drive the emitter of the following stage. By using emitter-follower stages as buffers in conjunction with the inter-stage coupling components, the necessary current amplification can be obtained.

Practical Bistable Circuit

The practical bistable circuit illustrated in Fig. 2 shows the addition of two transistors VT_3 and VT_4 used as emitter-followers, which enable the circuit to have the requisite loop gain to function as a bistable multivibrator. The hold-off resistors R_5 and R_6 in Fig. 1 can normally be omitted because the emitter potential of VT_3 is sufficient to hold off VT_2 when VT_1 is conducting, and similarly, the emitter potential of VT_4 is sufficient to hold off VT_1 when VT_2 is conducting. Capacitors C_1 and C_2 are included to improve the switching time.

Trigger pulses can be applied to the base electrodes of VT_1 and VT_2. Fig. 2 shows these electrodes strapped together and referred to ground potential by means of R_7 and the trigger pulses applied to the circuit via C_3. If, for

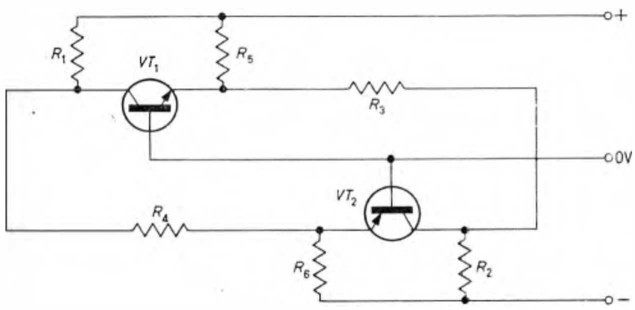
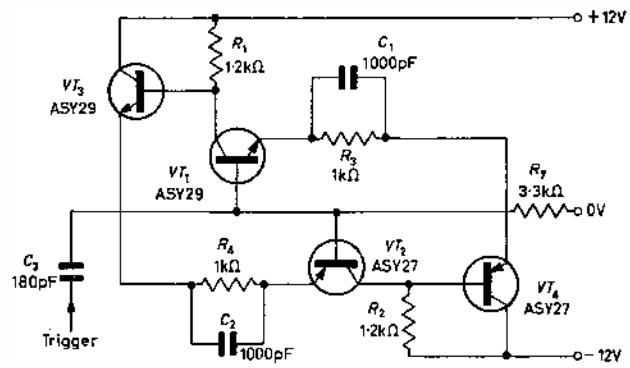


Fig. 1 (above). Basic bistable circuit

Fig. 2 (below). Practical bistable circuit



* *University of Nottingham*

example, the state when VT_1 is on and VT_2 is off is considered, then the circuit requires a negative going trigger pulse to change its state. By similar reasoning, a positive going trigger pulse is necessary for the circuit to revert to its original condition. This is a valuable property for applications where an indication of the change in the polarity of a pulse is required, such as in computers using positive and negative going pulses[1,2]. It is also convenient to use this circuit for pulse shaping, because it can be triggered on, say, by the differentiated leading edge of an incoming pulse and triggered off by its differentiated trailing edge.

Schmitt Trigger Circuit

The bistable circuit can easily be converted for Schmitt trigger operation by connecting the base electrodes of VT_1 and VT_2 to the input source via a resistor R_8 as shown in Fig. 3. For high speed operation, it may be found necessary to include a capacitor in parallel with R_8. The value of R_8 will depend on the impedance of the driving source and the permissible 'back-lash.'

A very valuable property of this circuit is that the change-over from one state to the other takes place when the input signal is varied through zero potential, the magnitude of the signal being low for either polarity. This is not easy to achieve with more conventional circuits without the use of backing-off potentials.

Monostable and Astable Circuits

It is possible to construct astable and monostable versions of the multivibrator, but the circuit must be modified so that the capacitive coupling is connected between the collector of the grounded-base stage and the next emitter-follower stage. It is now necessary to include a current limiting resistor in the collector circuit of the emitter-follower stage concerned. An astable circuit incorporating these modifications is shown in Fig. 4. The com-

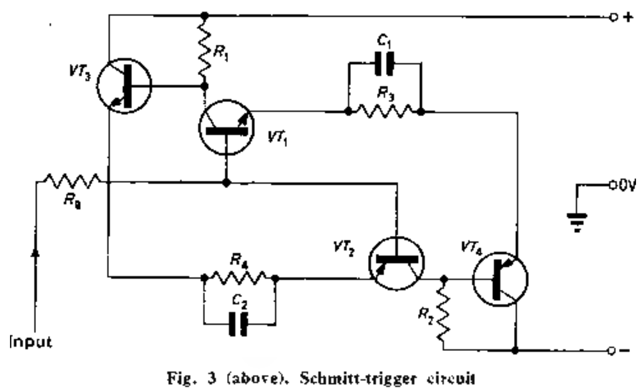
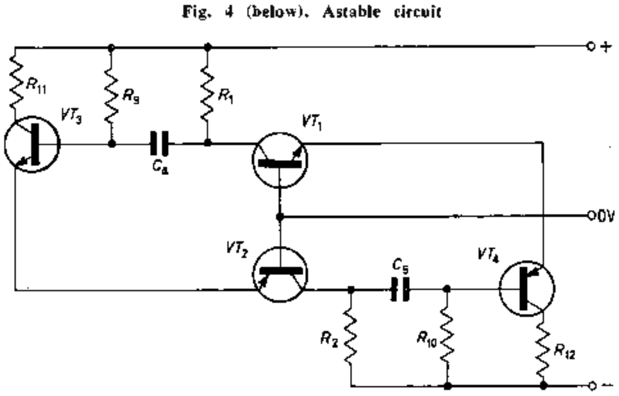


Fig. 3 (above). Schmitt-trigger circuit

Fig. 4 (below). Astable circuit



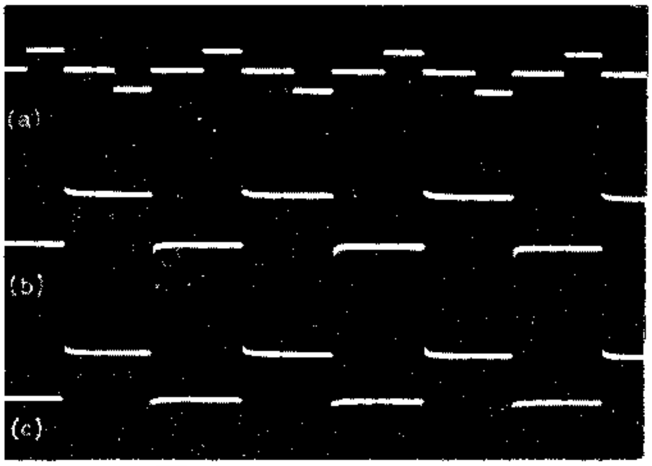


Fig. 5. Demonstration of bistable circuit. (Periodic time 100μsec)
(a) Input signal
(b) Waveform appearing at the collector of VT_1
(c) Waveform appearing at the collector of VT_2

ponents which determine the periodic time of oscillation are C_4 with R_9 and C_5 with R_{10}.

The monostable circuit would require only one of the interstage couplings of the astable version together with a direct coupling as used in the bistable version. Trigger pulses for the monostable version are applied to the base electrode of the appropriate grounded-base stage. The polarity of the trigger pulse is determined by the type of transistor used for this stage.

Conclusions

As an illustration, Fig. 5 shows the output waveforms obtained from the bistable circuit. Waveform (a) is the input signal to the circuit before differentiation. This sequence of positive and negative going pulses produces a differentiated pulse sequence of two successive positive 'spikes,' followed by two negative 'spikes.'

If the circuit is triggered, say, by the first positive edge then it will ignore the second positive edge, but will be triggered to its original state by the first negative edge. The second negative edge is again ignored. Waveforms (b) and (c) show the outputs from each collector of the grounded-base stages, which clearly demonstrates the sensitivity of the circuit to the polarity of the trigger pulses. The two outputs obtained are in the same phase, but waveform (b) is alternating between the positive rail potential and ground potential, and waveform (c) is alternating between ground potential and the negative rail potential.

This series of multivibrator circuits is a particularly useful addition to the range of circuits available for electronic digital systems which are symmetrical around ground potential, but its unique properties might find applications in many fields. Other examples of the novel circuits available when use is made of complementary pnp and npn devices have been given previously[3].

Acknowledgement

The Author would like to thank the University of Nottingham for providing the facilities to carry out this work.

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High Stability Low Frequency Band-Pass Filters

By P. G. Simpson*, B.Sc., A.M.I.E.E.

The design of very high stability v.l.f. band-pass filters, resulting in a phase stability of better than 1 part in 10^{10}, is described. Theoretical aspects of the choice of coupled circuits are discussed, and some accurate attenuation and phase angle curves are included, together with a brief summary of their derivation. Some practical design features of the completed filters are mentioned.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

A REQUIREMENT arose for the design of some very high phase-stability harmonic-selection filters. These have to select any one of a range of harmonically related sine waves from a rectangular wave pulse signal, and maintain any variation of phase due to LC product changes to less than 1 part in 10^{10} over a typical daily temperature range of 15°C. At the same time adequate rejection of unwanted harmonics is essential, and the whole equipment is to be kept to the minimum possible size, preferably within one chassis unit for each frequency range. The chassis are to be mounted in standard Post Office apparatus racks.

The article describes briefly the theoretical basis for the selection of the final design of coupled tuned circuits. Accurate attenuation and phase angle curves are required, and these are given, together with an outline of their derivation in the Appendix.

Coil and tuned circuit designs are outlined, including the reasons for component selection. A brief description is given of the insertion loss amplifiers. Practical design features are included, together with some performance figures.

Design Requirements

The main reason for the very high stability requirement of the filters is that the final equipment must have a frequency stability of better than 1 part in 10^{10} per day. Two filters are required, one for the range 1 to 2kc/s and the other for the range 14 to 18kc/s. As both of them contribute towards the final output, each one has to have a stability at least as good as 1 part in 10^{10} per day.

It can be seen that at f_o c/s, the total number of cycles per day C is

$$C = f_o \times 24 \times 3\,600 \text{ cycles}$$

A phase change θ is proportional to a change of cycles ΔC from

$$\theta/360 = \Delta C/1$$

Therefore as $\Delta C/C$ must be less than 10^{-10},

$$\Delta C/C = \theta/360 \times \frac{1}{f_o \times 24 \times 3\,600} \not> 10^{-10}$$

$$\therefore\ \theta \not> 3{\cdot}1 f_o \times 10^{-3}$$

$$\text{or } \theta \not> 3{\cdot}1 f \qquad (1)$$

where

f = frequency (kc/s)

θ = phase angle change per day (degrees)

For example, at the lowest frequency of 1kc/s, the phase angle change per day must be less than 3·1°, and at the highest frequency of 18kc/s, it must be less than 55·8°.

These changes occur as a result of an ambient temperature change which may reach 15°C over 24 hours. Therefore the resultant LC product change in the tuned circuits, or the equivalent resonant frequency change, must result in a phase change of less than that given in equation (1).

The filters are designed to select harmonics from rectangular pulses with pulse repetition frequencies of 100c/s (1 to 2kc/s harmonics) and 1 000c/s (14 to 18kc/s harmonics) and pulse widths of 100μsec and 10μsec respectively.

The specification calls for the adjacent harmonics to be rejected to greater than 60dB below the level of the wanted harmonic. In addition, the power output of the higher frequency filter is to be 10mW and that of the lower frequency filter is to be 0·1mW, both into 75Ω loads.

Attenuation and Phase Angle Curves

Investigation of the various types of filters soon showed that the best results could be obtained by using a series of coupled tuned circuits. A certain minimum degree of coupling is necessary for a given LC product or resonant frequency change in the tuned circuits in order to maintain the phase angle change within specification. This coupling factor would result in rejection figures of slightly more than half of the specified value in one pair of tuned circuits, assuming reasonable Q factors for the coils. Therefore it was found necessary to design two identical pairs of tuned circuits.

It was soon apparent from a study of the standard text books[1,2,3] that existing phase angle and attenuation curves neither cover the very small angles or give a sufficiently exact and comprehensive range of attenuation values. Further, it is known that attenuation curves are not symmetrical for low Q values and detailed information on this assymetry is not usually available.

Therefore a set of curves was prepared covering these points in some detail, and these are given together with a brief summary of the mathematical derivation. The curves were used for subsequent calculations in the design of the filters.

Temperature Coefficient Considerations

It was obvious that in addition to providing a very high Q, the inductors and capacitors would have to be matched as closely as possible in their temperature coefficients.

A comparison was made between ferrite cores and dust iron toroidal core coils, and it was concluded that the dust iron core coils were superior in terms of the required combination of high Q, minimum temperature coefficient variation and maximum operating power level. A similar survey of high stability capacitors showed that the best type to match the dust core figures was encapsulated polystyrene. In addition this type has very good figures for

* *Redifon Ltd.*

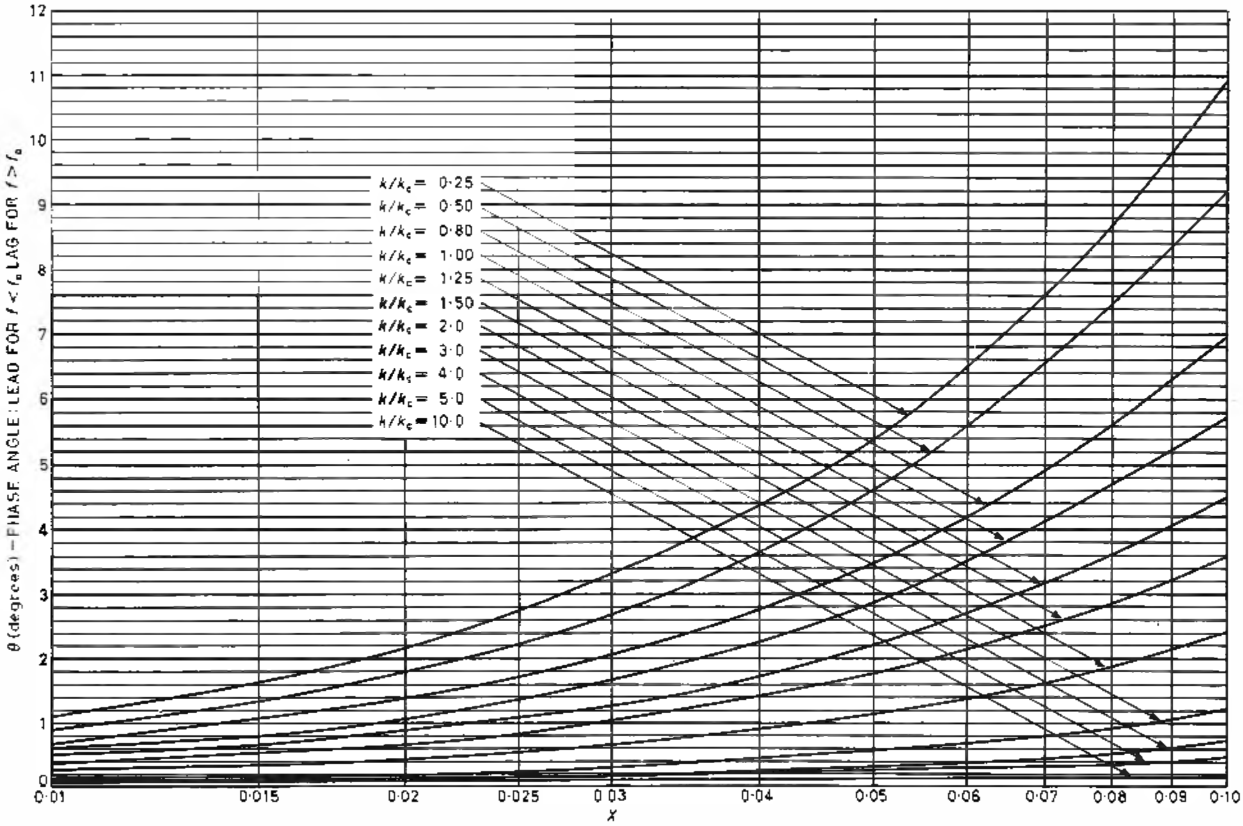


Fig. 1 (above). Universal phase angle curves : $Q \geq 20$ (Two coupled tuned circuits : $Q_p = Q_s$)

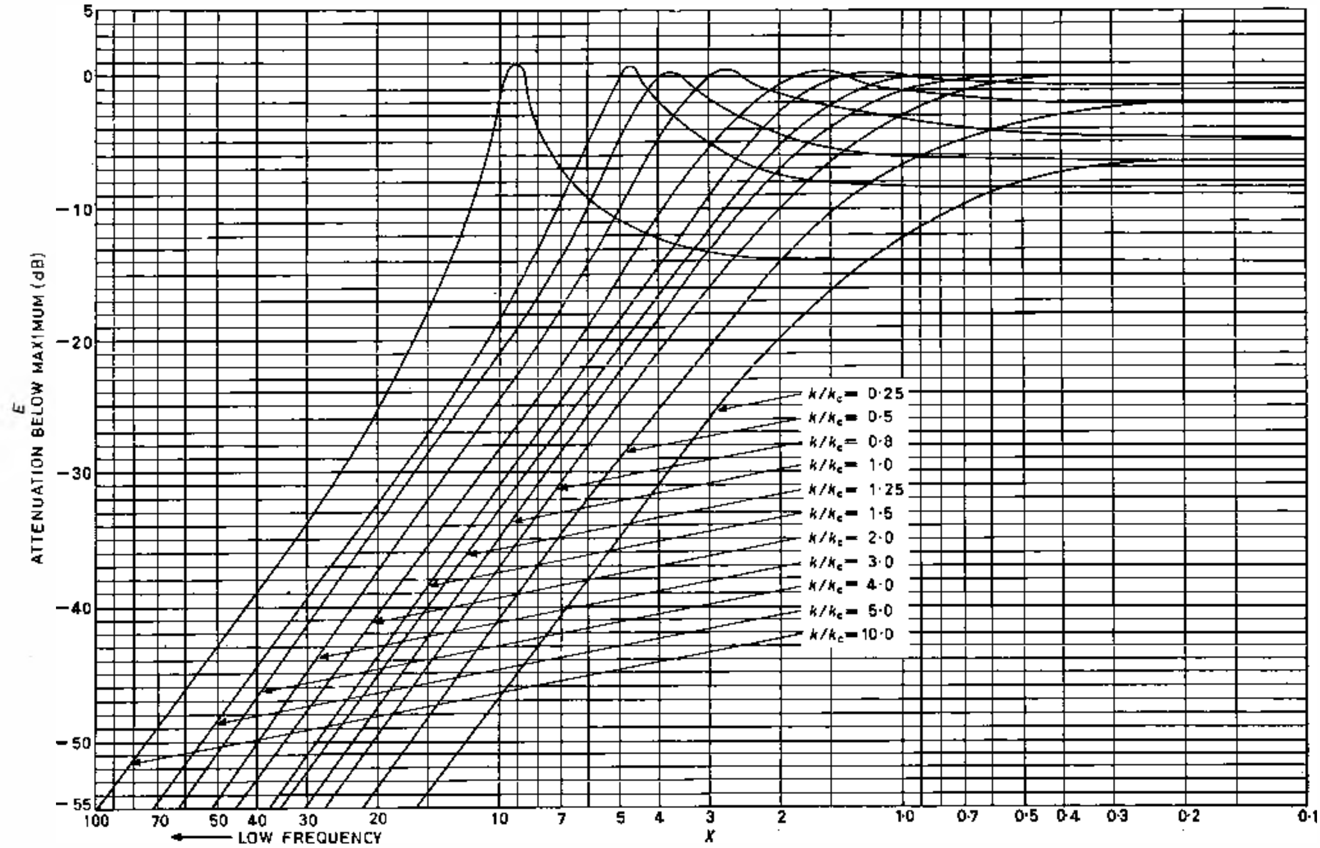
$$\theta = \tan^{-1}\left(\frac{2X}{1 + (k/kc)^2 - X^2}\right)$$

$X = 2Q(\gamma - 1)$, $\gamma = f/f_o > 1$ or $X = 2Q(1 - \gamma)$, $\gamma = f/f_o < 1$

Fig. 2 (below). Universal response curves : $Q \geq 100$ (Two coupled tuned circuits : $Q_p = Q_s$)

$X = 2Q(1-\gamma)$ $\gamma = f/f_o < 1$, $Q \geqslant 100$

$$E = -20\log \frac{Q^2\gamma^2}{2k/kc}\left[\left\{\frac{(k/kc)^2}{Q^2} + (1/Q^2) - (1-(1/\gamma^2))^2\right\} + (4/Q^2)(1-(1/\gamma^2))^2\right]^{\frac{1}{2}}$$



long term capacitance stability and is available with low volume/capacitance ratios. The cores and capacitors selected were 'Genelex' toroidal cores (Salford Electrical Instruments Ltd) and A597 polystyrene capacitors (G.E.C. Ltd.) respectively.

The dust-iron cores have a temperature coefficient of $+130 \pm 40 \times 10^{-6}/°C$, and the capacitors have a co-

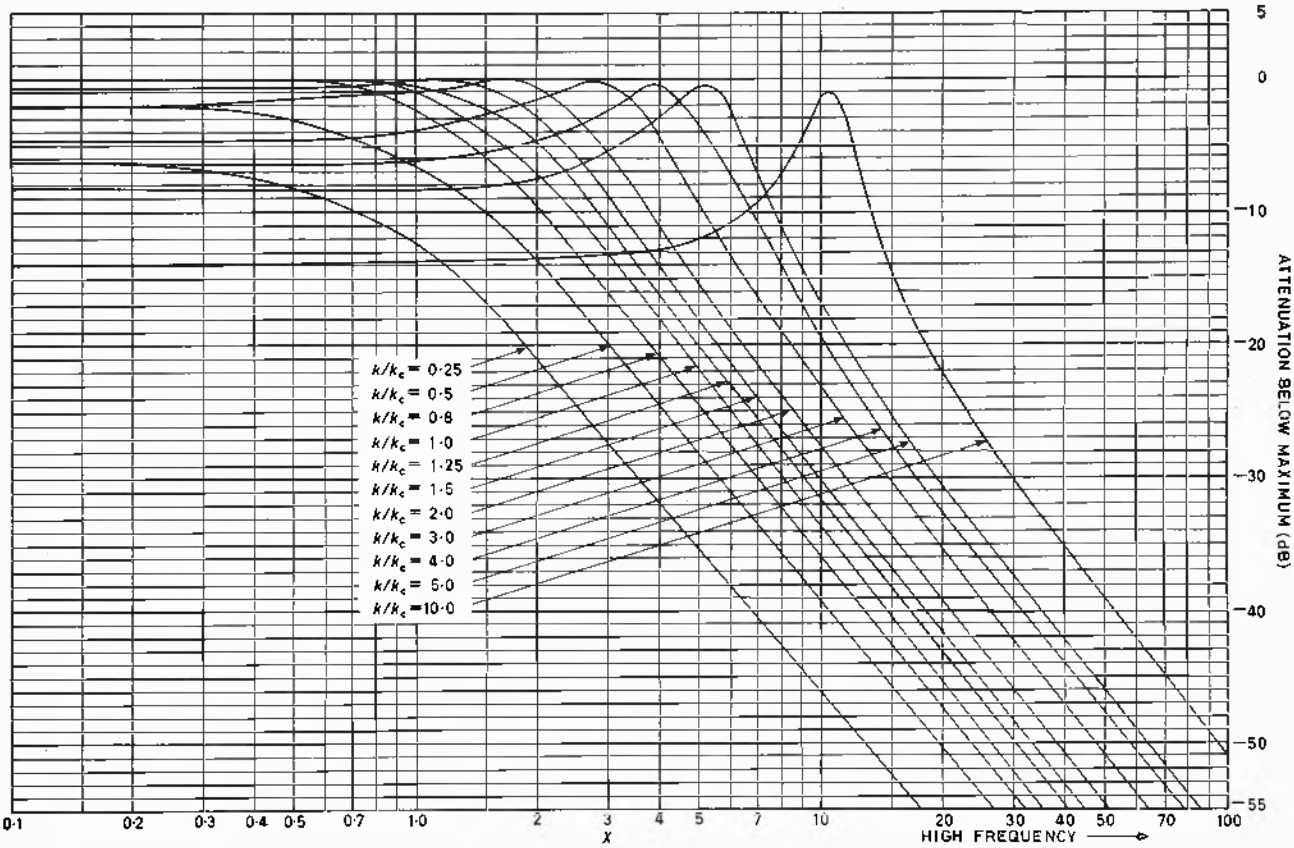


Fig. 3. Universal response curves: $Q \geq 100$
(Two coupled tuned circuits: $Q_p = Q_s$)

$X = 2Q(\gamma - 1)$ $\gamma = f/f_o > 1$, $Q \geqslant 100$

efficient of $-130 \pm 15 \times 10^{-6}/°C$. The maximum variation is therefore $\pm 55 \times 10^{-6}/°C$ or $\pm 825 \times 10^{-6}$ over the specified 15°C.

As the resonant frequency is proportional to the square root of the LC product, small fractional changes of resonant frequency are obtained by halving the fractional change of LC product. Therefore the maximum frequency change is $\pm 27{\cdot}5 \times 10^{-6}/°C$ or $\pm 412{\cdot}5 \times 10^{-6}$ over 15°C. For example, at 1kc/s the maximum frequency change is 0·413c/s.

Assuming for the moment that the self-Q of the coils can be taken as 200, and the coils are loaded to half this value, the effective Q will be 100. By definition the factor X is given from

$$X = 2Q(f/f_o - 1), \quad \text{where } f/f_o > 1$$

or

$$X = 2Q(1 - f/f_o), \quad \text{where } f/f_o < 1$$

The value of f differs from the resonant frequency f_o by f, where

$$X = 2Q(\Delta f/f_o)$$

The ratio $\Delta f/f_o$ over 15°C range is given above as $412{\cdot}5 \times 10^{-6}$.

Therefore

$$X = 2 \times 100 \times 412{\cdot}5 \times 10^{-6} = 0{\cdot}0825$$

Referring to Fig. 1, and taking 1kc/s as the example, i.e. a maximum phase angle change per day of 3·1°, the minimum coupling factor ratio k/k_c is just under 1·5. For 2kc/s, with a maximum phase angle change of 6·2°, the minimum k/k_c ratio is seen to be approximately 0·75.

Taking the above as a first approximation, the next step is to obtain the rejection of the 100c/s harmonics adjacent to 1kc/s and 2kc/s with the above k/k_c ratios. Here the values of X are:

$$X = 2 \times 100\,((1100/1000) - 1) = 20$$

and

$$X = 2 \times 100 \times (1 - (900/1000)) = 20$$

for 1kc/s, and

$$X = 2 \times 100\,((2100/2000) - 1) = 10$$

and

$$X = 2 \times 100\,(1 - (1900/2000)) = 10$$

for 2kc/s.

From Fig. 2, the lower harmonic rejection for 1kc/s at $k/k_o = 1{\cdot}5$ is 43·5dB, and for 2kc/s at $k/k_c = 0{\cdot}75$ it is 38dB. Fig. 3 shows that the corresponding figures for the upper harmonics are 41·5 and 36·5dB.

The above calculations confirm that one pair of tuned circuits will not give the specified rejection (over 60dB) and maintain a phase angle change of less than the amount required to meet the daily stability figure. They also show that two pairs of tuned circuits are sufficient (e.g. $2 \times 41{\cdot}5 =$ 83dB at 1kc/s, and $2 \times 36{\cdot}5 = 73$dB at 2kc/s).

Coil Design

In order to achieve coils with self-Q values greater than 200 over the two frequency ranges required, consideration was given to the value of core permeability and the type of winding, as well as coil finishing treatments.

The lower frequencies indicated a high permeability and a value of 140 was selected. The higher frequencies were covered by a permeability of 25.

The values of inductance were determined by the available capacitors and the lowest frequency in each range. It is desirable to keep the inductance value low in order to match into the buffer amplifiers. The highest capacitance value in the A597 range is 0·51μF and a convenient inductance value to tune this to 1kc/s is 50mH, and for 14kc/s it is 6mH.

Coils were wound for the higher frequency range using both 81/42 Litz wire and 18 s.w.g. enamelled copper wire. Measurements indicated that over 14 to 18kc/s Q values over 300 could be obtained with the Litz wire coil compared to 200 with the single strand copper wire, and therefore the Litz wire designs were selected.

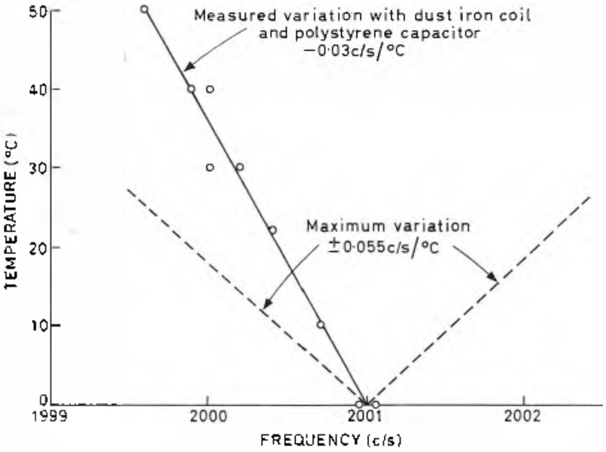


Fig. 4. Resonant frequency-temperature curves

The self-capacitance was found to be rather higher than expected, and varied between 100 and 200pF.

Various impregnants were tried including moulded Araldite resins AY18/HZ18, MY750/HT972 and 820 as well as wax. The resins increased the self-capacitance by up to three times the untreated values, and the wax increased the values by a factor of about 1·5. The 820 resin decreased the Litz wire Q values to 25 per cent of the original values. The other impregnants made no difference to the Q values, but careful measurements of inductance over a long period of time showed that this increased slightly, possibly due to changes in self-capacitance as well as winding variations. It was finally decided to leave the coils untreated.

Measurements were also taken on the effect of voltage on the self-Q of the toroidal coils and as expected in all cases increase of voltage levels resulted in an increase of hysteresis loss and decrease of Q value. A value of 4V was selected as the Q figures are still high at this level, and this value is also suitable for the insertion loss amplifiers, which are designed for use with a 12V d.c. supply.

Tuned Circuits and Coupling Methods

Some consideration was given to methods of coupling, including top and bottom capacitance and inductance coupling, and resistive coupling. The method selected was mutual inductance coupling as this does not require extra components as used in other methods, and it is fairly simple to obtain the correct degree of coupling required.

From the phase angle-resonant frequency curves in Appendix (1), it is a simple matter to select the minimum k/k_c ratio, given the frequency and the effective Q value together with the maximum allowable phase angle. This ratio will then determine the rejection for the adjacent harmonics in each case. The coupling factor itself is derived from the known value of k/k_c and the definition of k_c as $1/Q$, which is also known. Therefore as k/k_c is also the ratio of coupling turns to total turns, the coupling turns are obtainable, given the total turns.

Amplifier Design

The insertion loss of the first pair of coupled tuned circuits is compensated by the inclusion of a simple two-stage transistor buffer amplifier of the Bootstrap type[4]. This has the advantage that its output impedance is high compared to that of a conventional single-stage circuit,

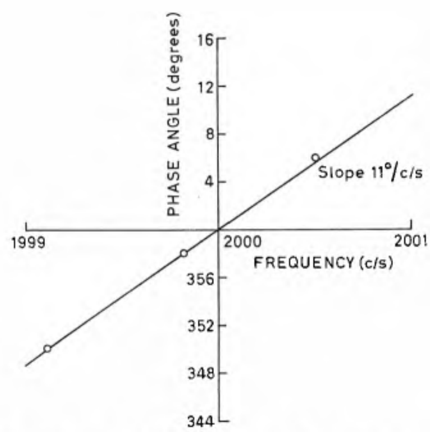


Fig. 5. Phase angle-frequency curve 2kc/s filter

and it can therefore be tapped into the following coil at a convenient point. The effective parallel impedance of the tuned circuit pair although kept to a minimum, is of the order of several hundred kilohms, and this requires a high source impedance to prevent over damping.

In the high frequency filter, a second amplifier is used after the second pair of tuned circuits as the output requirement is 10mW.

Silicon transistors are used where possible to improve temperature-gain stability.

Practical Filter Requirements

As 11 and 5 different frequencies are required in the low and high frequency range respectively, one common coil is used for each range, and the appropriate capacitor is switched into the circuit. This means that each input and output tap as well as each set of coupling turns has to be switched for the required frequency. This is accomplished automatically by the use of Ledex relay switches, remotely controlled.

Typical Performance Results

Fig. 4 shows the variations of resonant frequency plotted against temperature for a 2kc/s filter. The resonant frequency at each temperature is obtained by determining very accurately the small phase angle changes resulting from injecting signals either side of resonance[5]. These are plotted against frequency and resonance is obtained where the line crosses the frequency axis.

As seen earlier, the temperature coefficient can change by $\pm 27{\cdot}5 \times 10^{-6}/^{\circ}C$. At 2 000c/s this is $\pm 0{\cdot}055$c/s/°C and this is plotted, together with the measured change. It is obvious that the measured change is well within the maximum and represents a value of $0{\cdot}03 \times 15 = 0{\cdot}45$c/s for 15°C, the specified ambient variation.

The method of phase difference measurement is also used to show the slope of phase against resonant frequency, and this is seen in Fig. 5 for the 2kc/s filter. Here the phase angle changes by 11° for 1c/s shift. As the actual resonant change is measured as 0·45c/s the corresponding phase change would be $0{\cdot}45 \times 11 = 5°$. From equation (1) the maximum allowable change for 2 000c/s is 6·2°, and therefore the change of LC product or resonant frequency over 15°C is within the maximum permissible value.

This same filter results in adjacent harmonic rejection figures of over 70dB either side of resonance.

Acknowledgment

The author wishes to thank Redifon Ltd. Wandsworth, London, England, and Continental Electronics Systems Inc, Dallas, Texas, U.S.A., for whom the work was carried out, for permission to publish this article.

Appendix

(1) Accurate Universal Response and Phase Angle Curves

Universal response or resonance curves and phase angle curves for two coupled tuned circuits are available in most standard reference books. However, these are not usually sufficiently accurate to enable results to be obtained to the nearest decibel of response at different frequencies or to the nearest degree of phase change. In addition the curves are given for only a few discrete values of coupling factor ratio (k/k_c), making it difficult to interpolate accurately for other values. While a considerable amount of design work can be derived from these curves to a first order approximation, occasionally the need arises for more acurate and exact information.

The same results can be obtained by using accurate design formulae but this is often a long and tedious process, especially where several trial calculations have to be made. Therefore the following curves have been carefully drawn based on the design formulae, and they provide a rapid method of design for a wide variety of requirements.

Universal Response Curves –Derivation

For definition of subscripts 1 and 2, see end of this section.

The basic formulae for attenuation is

$$E = -20 \log E_m/E_f \text{ dB} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

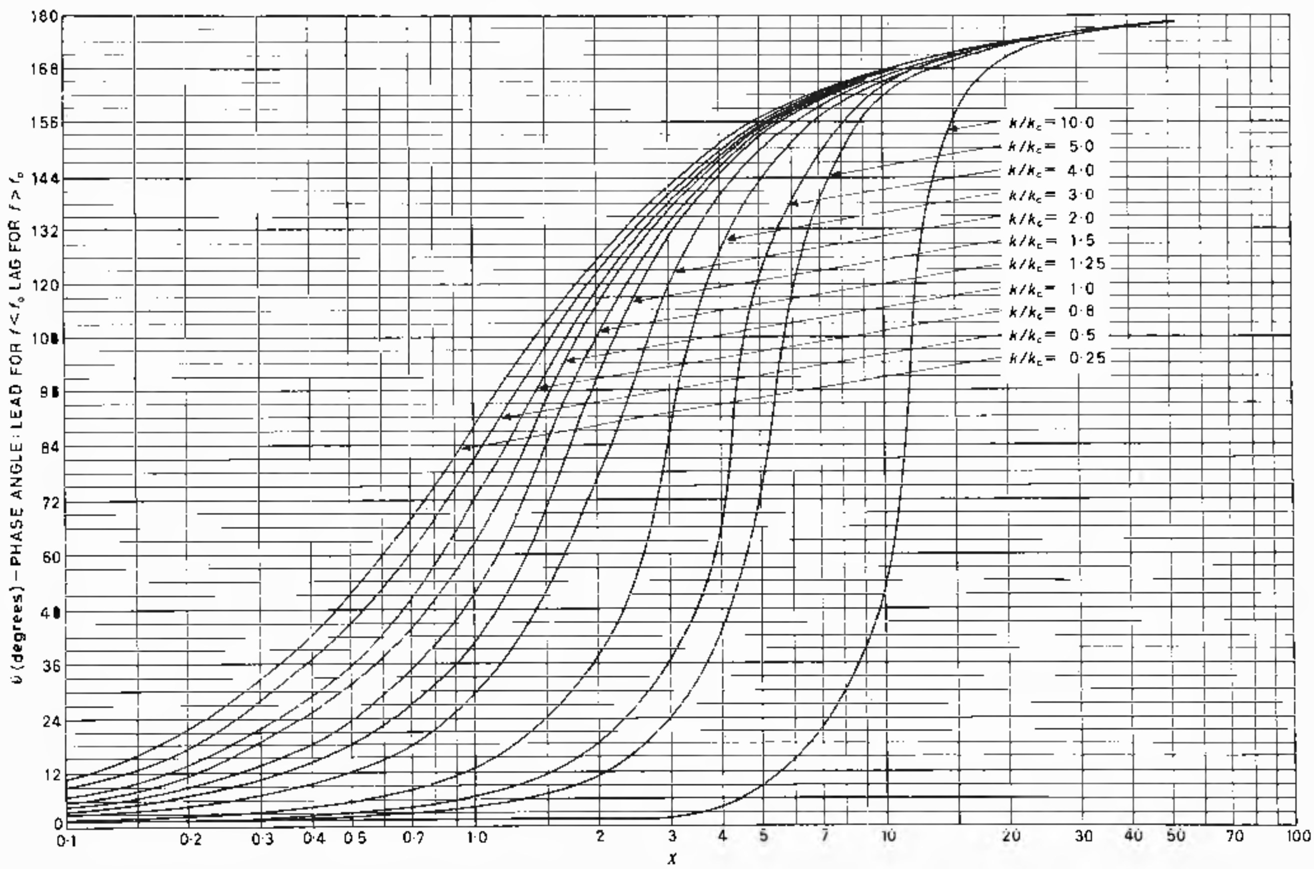
where

$$E_m/E_f = \frac{\text{Maximum possible secondary voltage at optimum coupling}}{\text{Secondary voltage at frequency } f}$$

Fig. 6. Universal phase angle curves: $Q \geq 100$
(Two coupled tuned circuits: $Q_p = Q_s$)

$$X = 2Q\,(\gamma = f/f_o > 1 \text{ or } X = 2Q\,(1-\gamma),\ \gamma f/f_o < 1$$

$$\theta = \tan^{-1} \frac{2Q\,(1-(1/\gamma^2))}{(k/kc)^2 + 1Q^2\,(1-(1/\gamma^2))^2}$$



The ratio is given from

$E = -20 \log E_m/E_f$

$$= -20 \log \frac{Q^2 \gamma^2}{2\,k/k_c} \{[(k/k_c)^2\, 1/Q^2 + 1/Q^2 - (1-(1/\gamma^2))^2]^2 + 4/Q^2 (1-(1/\gamma^2))^2\}^{\frac{1}{2}} \quad \ldots\ldots\ldots (3)$$

where

Q = Primary and secondary Q (assumed equal)

f = frequency of applied signal

f_o = resonant frequency of primary and secondary circuits

$\gamma = f/f_o$

k = coupling factor

k_c = critical coupling ($= 1/Q$)

This formula is a simplification of that given in Terman's 'Radio Engineering Handbook', page 159 (1943 edition), for circuits of equal primary and secondary Q values.

At resonance, $\gamma = 1$ and E_m/E_{fo} reduces to

$$E_m/E_{fo} = \frac{1 + (k/k_c)^2}{2\,k/k_c} \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

By definition, the peak-to-trough attenuation A is

$$A = -20 \log E_m/E_o \quad \ldots\ldots\ldots\ldots\ldots (5)$$

As a result, the attenuation below resonance is

$$A_T = E - A \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

The peak frequency is given by

$$\gamma_p = \frac{1}{\sqrt{\{1 \pm k/k_c (1/Q)\,[1-(1/(k/k_c)^2)]^{\frac{1}{2}}\}}} \quad \ldots\ldots (7)$$

where

$\gamma_p = f_p/f_o$

f_p = peak frequency

At high values of Q the curves are all symmetrical about f_o, but when $Q < 200$, they become unsymmetrical. Therefore the attenuation is different below and above resonance and the subscripts 1 and 2 are used to denote frequencies less than or greater than resonance respectively.

The peak-to-trough attenuations are therefore

$$A_1 = |A \sim E_{p1}|\ \text{dB} \quad \ldots\ldots\ldots\ldots\ldots (8)$$

and

$$A_2 = |A \sim E_{p2}|\ \text{dB} \quad \ldots\ldots\ldots\ldots\ldots (9)$$

Values of E_{p1} and E_{p2} are determined by substituting equation (8) in equation (4). This reduces to

$$E_p = -20 \log \gamma_p^2$$

Utilization of Response Curves

Figs. 2 and 3 consist of the response curves derived from equations (3), (5) and (7). They are plotted in terms of X where

$$X = 2Q\,(1 - \gamma) \quad \text{for } \gamma = f/f_o < 1,\ \ Q \geqslant 100 \quad \ldots\ldots\ldots\ldots (10)$$

or

$$X = 2Q\,(\gamma - 1) \quad \text{for } \gamma = f/f_o > 1,\ \ Q \geqslant 100 \quad \ldots\ldots\ldots\ldots (11)$$

This is a convenient combination for difference values of Q and off-resonant frequencies.

The range and number of k/k_c values should cover nearly all requirements, and interpolations can be made to the nearest decibel in most cases.

If Q is large such that X is off the X-axis of the curves then equation (3) can be simplified to

$$E = -20 \log \frac{Q^2 \gamma^2}{2\,k/k_c} [(k/k_c)^2 (1/Q^2) - (1-(1/\gamma^2))^2] \quad \ldots\ldots\ldots\ldots (12)$$

This enables the attenuation below resonant value to be calculated, as the same peak-to-trough attenuation holds for each k/k_c curves, irrespective of the X values.

It can be seen from the curves that they are assymmetrical by 1 to 2dB at frequencies equally below and above resonance.

Universal Phase Angle Curves—Derivation

The ratio of maximum secondary voltage to secondary voltage at frequency f was originally derived from

$$E_m/E_f = \frac{Q^2 \gamma^2}{2\,k/k_c} \{[(k/k_c)^2\, 1/Q^2 + 1/Q^2 - (1-(1/\gamma^2))^2] - j(2/Q)\,(1-(1/\gamma^2))\} \quad \ldots\ldots\ldots\ldots (13)$$

This is equivalent to

$$E_m/E_f = K\,[R + jX] \quad \ldots\ldots\ldots\ldots (14)$$

Therefore the phase angle is given by

$$\theta = \tan^{-1} X/R = \tan^{-1} \frac{2Q\,(1-(1/\gamma^2))}{(k/k_c)^2 + 1 - Q^2\,(1-(1/\gamma^2))^2} \quad \ldots\ldots\ldots\ldots (15)$$

Curves of this equation are plotted in Fig. 6 for values of X, given by equations (10) and (11).

Where $\gamma \to 1$, i.e. the signal frequency is very close to the resonant frequency, and X becomes small

$$(1 - (1/\gamma^2)) \to 2\,(\gamma - 1)$$

Therefore

$$\theta = \tan^{-1} \frac{4Q\,(\gamma - 1)}{(k/k_c)^2 + 1 - 4Q^2\,(\gamma - 1)^2} \quad \ldots\ldots (16)$$

or

$$\theta = \tan^{-1} \frac{2X}{(k/k_c)^2 + 1 - X^2} \quad \ldots\ldots\ldots\ldots\ldots (17)$$

Curves of equation (17) are plotted in Fig. 1 for small values of X between 0·01 and 0·1.

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A Stepping Drum Programmer

A stepping drum programmer is now being manufactured by the Woden Transformer Co. Ltd., of Bilston, Staffordshire, under licence from the Tenor Company, of Wisconsin, U.S.A.

The programmer provides a basic method for controlling load circuits in an interlocked predetermined sequence.

Load circuit switches are actuated by nylon plugs placed in discrete locations on the surface of the memory drum. New combinations of plugs are positioned to actuate the row of load circuit switches as the memory drum is stepped.

The memory drum is stepped by the closure of a selected (external) condition-sensing switch. Typical condition-sensing switches include timers, counters, push-buttons, limit switches, temperature switches, etc.

Features of this equipment are: rapid programme changes without rewiring; logic and circuit design costs are reduced by inclusion of interlocking relay logic; models are available to accommodate 31, 62 or 93 separate load switches for 30, 60 and 100 stages; it will switch up to 10A loads. A single drum unit will mount in a standard rack.

Two Distinct Boundaries for Feedback Transistor Oscillators

By B. Shahzadi*

Two different criteria for sustained oscillation can be found in the literature for feedback transistor oscillators. Authors usually indicate one of the criteria and ignore the other one inadvertently. In this article the criteria are worked out and discussed for several types of feedback oscillators.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

THE criterion for and the frequency of oscillation of a feedback oscillator may be determined by several methods. One method, and perhaps the easiest, is to set the real and imaginary parts of the equivalent circuit determinant equal to zero[1-3]. An alternate procedure is to write the required equations in order that the output current and voltage of the feedback network shall be equal in phase and magnitude to the input current and voltage of the amplifier[1]. Analysis by differential equations is another way of attacking the problem.[4] Certainly the result will be the same regardless of the applied method.

In the case of a valve oscillator, the analysis is fairly simple and the literature reveals a unique answer for the condition for oscillation. For a transistor oscillator, however, the circuit is more complicated, and it gives rise to two different criteria for oscillation.

Some authors[1,2,3,5] obtain the criterion for the Colpitts oscillator of Fig. 1 as

$$C_2/C_1 \simeq \Delta_{he}/h_{fe} \simeq \Delta_{hb} \quad \ldots\ldots\ldots\ldots (1)$$

or equivalent expressions, where

$$\Delta_{he} = h_{ie}h_{oe} - h_{re}h_{fe}$$
$$\Delta_{hb} = h_{ib}h_{ob} - h_{rb}h_{fb}$$

Other authors[4,6,7] end up with the criterion:

$$C_2/C_1 \simeq h_{fe} \quad \ldots\ldots\ldots\ldots (2)$$

Obviously equations (1) and (2) are two completely different results. Or in other words, they seem to be in contradiction, since usually $\Delta_{hb} \ll 1$ and $h_{fe} \gg 1$.

The objective here is to find the exact solution and clear up the above confusion which, as will be seen, arises from some over-simplifications. In fact, it will be shown that either of the answers is correct, and the oscillatory condition lies between the two boundaries determined by equations (1) and (2).

Analysis of Feedback Oscillators

In this section several of the most usual types of feedback oscillators will be analysed by setting the network determinant equal to zero.

COLPITTS OSCILLATOR

The a.c. equivalent circuit of the Colpitts oscillator of Fig. 1, using common-emitter hybrid parameters, is shown in Fig. 2 in appropriate form for node analysis.

Neglecting the coil resistance, the node admittance matrix of the equivalent circuit is

$$Y = \begin{bmatrix} sC_1+(1/sL)+(1/h_{ie}) & -((1/sL)+(h_{re}/h_{ie})) \\ (h_{fe}/h_{ie})-(1/sL) & (1/sL)+sC_2+(\Delta_{he}/h_{ie}) \end{bmatrix} \quad \ldots\ldots\ldots\ldots (3)$$

Letting $s = j\omega$ and equating the real part of Y to zero, the frequency of oscillation is obtained:

$$\omega^2 = \frac{C_1 + C_2}{LC_1C_2} + \frac{h_{oe}}{h_{ie}C_1C_2} \quad \ldots\ldots\ldots\ldots (4)$$

Introducing

$$(1/C) = (1/C_1) + (1/C_2)$$

equation (4) becomes

$$\omega^2 = (1/LC) + \frac{h_{oe}}{h_{ie}C_1C_2} \quad \ldots\ldots\ldots\ldots (5)$$

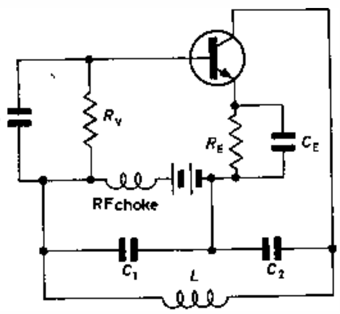

Fig. 1. Colpitts oscillator

The second term in equation (5) has a small effect on ω, for practical values of the parameters. Therefore, the frequency of oscillation depends mainly on the resonant elements C_1, C_2 and L.

Assuming $\omega^2 = 1/LC$ and setting the imaginary part of Y equal to zero, the following equation is obtained.

$$(C_2/C_1)^2 - (h_{fe} - h_{re})(C_2/C_1) + \Delta_{he} = 0 \quad \ldots\ldots (6)$$

Equation (6) is in a quadratic form with respect to C_2/C_1 and, in general, there should be two roots for the equation. But it is easy to miss one of the roots, as has happened, by over-simplification.

In order to overcome this difficulty and obtain both of the solutions, it is helpful to replace the hybrid parameters by some typical values and then investigate the nature of the two roots of equation (6). Using the typical values:

$$h_{ie} = 2\text{k}\Omega$$
$$h_{re} = 5 \times 10^{-4}$$
$$h_{fe} = 50$$
$$h_{oe} = 25 \times 10^{-6}\Omega^{-1}$$

equation (6) becomes approximately

$$x^2 - 50x + 0{\cdot}025 = 0 \quad \ldots\ldots\ldots\ldots (7)$$

where $x = C_2/C_1$

It is seen that the product of the two roots, i.e.

$$x_1x_2 = 0{\cdot}025$$

is much smaller than unity, while the sum of the roots, $x_1 + x_2 = 50$, is much larger than unity. This may happen only when one of the two positive roots is much smaller and the other much larger than unity. Therefore, the two roots can be calculated by assuming, first x is very

* *The University of Wisconsin, U.S.A.*

small and next very large in equation (7). For very small x, equation (7) reduces approximately to

$$-50x + 0{\cdot}025 = 0$$

which gives the root

$$x = C_2/C_1 = 1/2000.$$

For very large x, equation (7) can be written approximately as

$$x^2 - 50x = 0$$

or

$$x = C_2/C_1 = 50$$

Referring back to equation (6) and neglecting h_{re} in comparison to h_{ie}, for $C_2/C_1 \ll 1$,

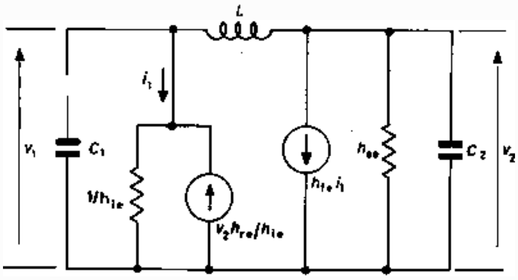

Fig. 2. Common-emitter equivalent circuit of the Colpitts oscillator of Fig. 1.

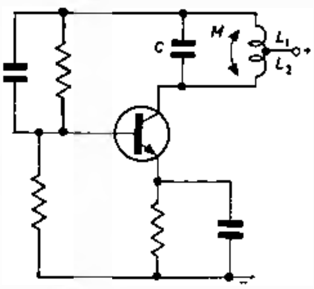

Fig. 3. Hartley oscillator

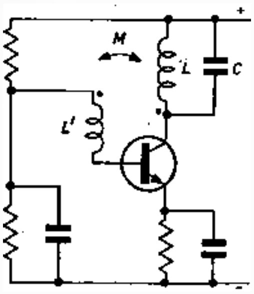

Fig. 4. Tuned-collector oscillator

$$C_2/C_1 \simeq \Delta_{he}/h_{fe} \quad \ldots\ldots\ldots\ldots\ldots (8)$$

And for $C_2C_1 \gg 1$, the equation gives

$$C_2/C_1 \simeq h_{fe} \quad \ldots\ldots\ldots\ldots\ldots (9)$$

Equations (8) and (9) give two different boundaries for oscillation. Some simple reasoning shows that the oscillator will oscillate if

$$\Delta_{he}/h_{fe} \leqslant C_2/C_1 \leqslant h_{fe} \quad \ldots\ldots\ldots\ldots (10)$$

Hartley and Tuned-type Oscillators

Using the same procedure as above, the condition for sustained oscillation for the Hartley oscillator of Fig. 3 is found as

$$1/h_{fe} \leqslant \frac{L_2 + M}{L_1 + M} \leqslant h_{fe}/\Delta_{he} \quad \ldots\ldots\ldots\ldots (11)$$

For the tuned-collector (tuned-output) oscillator of Fig. 4, the analysis results in

$$1/h_{fe} \leqslant L/M \leqslant h_{fe}/\Delta_{he} \quad \ldots\ldots\ldots\ldots (12)$$

And for the tuned-base (tuned-input) oscillator shown in Fig. 5,

$$1/h_{fe} \leqslant M/L \leqslant h_{fe}/\Delta_{he} \quad \ldots\ldots\ldots\ldots (13)$$

In order to generalize the solutions, first it must be noted that by writing expression (10) as

$$1/h_{fe} \leqslant C_1/C_2 \leqslant h_{fe}/\Delta_{he} \quad \ldots\ldots\ldots\ldots (14)$$

it is in the same form as expressions (11), (12) and (13); which may be written in a general form, namely

$$1/h_{fe} \leqslant Z_{out}/Z_{in} \leqslant h_{fe}/\Delta_{he} \quad \ldots\ldots\ldots\ldots (15)$$

where

for a Colpitts oscillator $Z_{out} = 1/sC_2$ and $Z_{in} = 1/sC_1$

for a Hartley oscillator $Z_{out} = s(L_2+M)$ and $Z_{in} = s(L_1+M)$

for a tuned-collector oscillator $Z_{out} = sL$ and $Z_{in} = sM$

for a tuned-base oscillator $Z_{out} = sM$ and $Z_{in} = sL$

Next, it is clear that in any types of the above oscillators, the boundaries of oscillation for common-base and common-collector configurations can be obtained by replacing the common-emitter h-parameters of expression (15) by appropriate h-parameters. But since h_{fb} and h_{fc} are negative, a general form of equation (15) which holds for all three configurations would be

$$1/|h_f| \leqslant |Z_{out}/Z_{in}| \leqslant |h_f|/\Delta_h \quad \ldots\ldots\ldots (16)$$

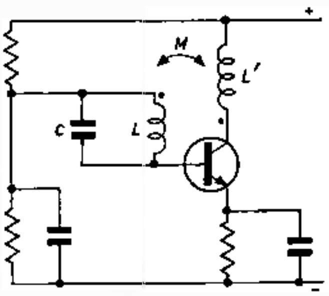

Fig. 5. Tuned-base oscillator

The negativeness of h_{fb} and h_{fc} implies, however, that for common base and common collector configurations, Z_{out} and Z_{in} must have opposite signs. The implication is that for tuned-type oscillators, M must be negative; which means one of the coils should be reversed compared to the polarity shown in Figs. 4 and 5. Furthermore, there cannot be oscillation in Colpitts and Hartley type circuits in common base and common collector configurations. In other words, the tank circuit in Colpitts and Hartley oscillators can be placed only across the base and collector and not elsewhere.

Experimental Results

In order to investigate the validity of the above theory, the circuit of Fig. 1 was tested with the following numerical values:

$R_T = 100\text{k}\Omega$

$R_E = 4\text{k}\Omega$

$C_T = C_E = 0{\cdot}01\mu\text{F}$

$L = 10\text{mH}$ (with a very high selectivity factor, $Q = 320$)

$V_{CC} = 6\text{V}$

which produced an emitter current of about 1mA and a collector to emitter voltage of about 3V. The transistor used in the circuit was a 2N338 silicon npn type with the following values for hybrid parameters under the above conditions and at room temperature.

$h_{ie} = 2200\Omega$

$h_{re} = 4 \times 10^{-4}$

$h_{fe} = 80$

$h_{oe} = 17\mu\Omega^{-1}$

$\Delta_{he} = 5{\cdot}4 \times 10^{-3}$

Two sets of values for C_1 and C_2 at the critical points of oscillation were:

(1) $C_1 \simeq 0{\cdot}2\mu F$

$C_2 \simeq 50pF$

with the ratio

$C_1/C_2 = 4000$

(2) $C_1 \simeq 30pF$

$C_2 \simeq 2100pF$

with

$C_1/C_2 = 1/70$

And the oscillator operated if

$$1/70 \leqslant C_1/C_2 \leqslant 4000$$

Placing the above set of h-parameters in expression (14), the theoretical condition for oscillation is obtained as

$$1/80 \leqslant C_1/C_2 \leqslant 14800$$

The result would seem satisfactory considering the approximations made in deriving the theoretical formulas, and the fact that the transistor parameters vary widely from one transistor to another, even of the same type.

Conclusion

A simple analysis of feedback transistor oscillators has been presented to determine the conditions for oscillation. It was found that there are two boundaries of oscillation and the oscillator is active if feedback elements satisfy the condition of expression (16) for Colpitts, Hartley and tuned-type oscillators. Thus there are two choices for indicating the boundary of oscillation, and although either of them can be considered in the design of an oscillator, one of the two might be advantageous in some respects, in any particular situation.

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A Combined Voltage and Earth Indicator for D.C. Supplies

By G. T. Edwards*

A simple circuit is described for use with d.c. supply units; it indicates the output and which, if any, of the output terminals is earth. It is suitable for use where one supply unit feeds a number of remote outlets.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

THE development and testing of electronic circuits and systems often involves the use of more than one direct voltage supply unit, each connected to give one of three conditions:

(1) Negative output terminal earthed.

(2) Positive output terminal earthed.

(3) 'Floating' Supply (i.e. neither output terminal earthed.)

A simple metering system is described here which helps to avoid confusion and generally ease the understanding of d.c. circuit conditions, particularly when operating with a number of supply rails of various voltages and polarities.

Basic System

A centre-zero meter (M) is connected as shown in Fig. 1 so that it can display two pieces of information:

(a) The output voltage

(b) Which, if any, of the output terminals is earthed.

R_1 and R_2 are equal in value, one of these two resistors serving as a voltage multiplier for the meter when an output terminal is earthed. The other resistor, being many times greater than the internal resistance of the meter movement (M), has a negligible shunting effect. Calculation of the values for R_1 and R_3 is made using the familiar formula for voltage multiplier resistors:

$$R_1 = R_2 = (V/I_m) - R_m$$

where

V = required full-scale meter reading in volts

I_m = current required by the meter for full-scale deflexion in amperes

R_m = internal resistance of the meter in ohms.

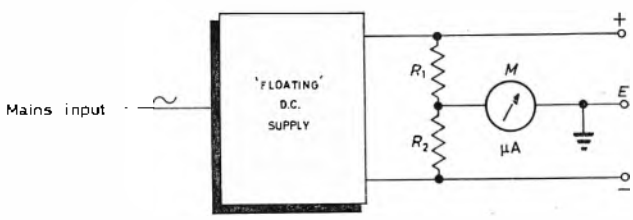
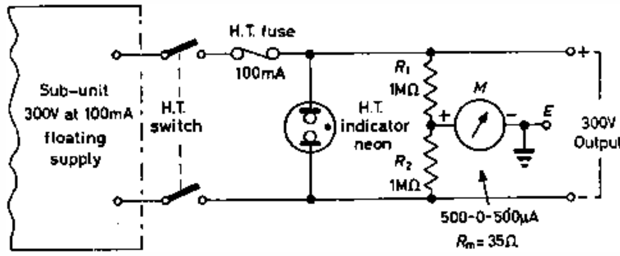


Fig. 1 (above). Basic system

Fig. 2 (below). Typical output circuit details



* *Formerly Atomic Weapons Research Establishment.*

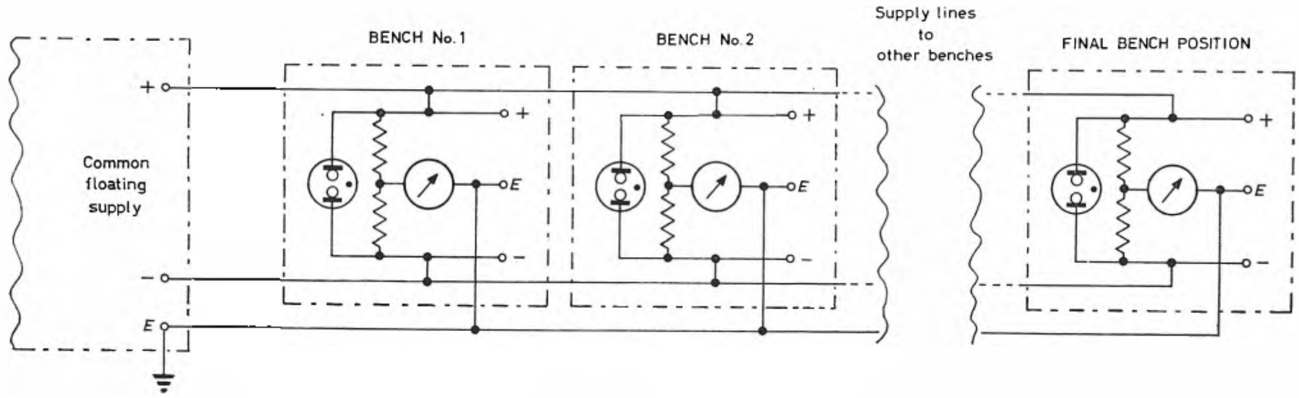


Fig. 3. Basic circuit for individual bench indicators

A Typical Unit

An example of this system is shown in Fig. 2. As the direction of current flow through the meter is dependent upon which output terminal is earthed one half of the scale may be marked 'positive earthed' and the other 'negative earthed'.

In cases where neither side is earthed there will be zero indication on the meter. Even though the meter indicates zero voltage in this condition, ambiguity is avoided by the provision of a neon bulb as a separate indicator to show that in fact an output voltage is present. A convenient type of neon is the Arcolectric SL 162 which includes a limiting resistor not shown in the diagrams.

For fixed low voltage supplies an electromechanical indicator, such as the Autophone Star Indicator type 7182, would be a suitable alternative to a neon bulb.

Variable Voltage Supplies

While this metering system is best applied to fixed voltage supply units so that imperfect earthing or unwanted leakages are indicated by deflexions that do not fall on marked anticipated points on the scale, there is no reason why it should not be usefully applied to variable output units. For such units it is recommended that a separate conventional voltmeter be included (instead of the neon bulb in Fig. 3) to indicate magnitude of the output voltage, and an inexpensive unscaled centre-zero meter included (with the associated resistors R_1 and R_2) to indicate the earth connexion.

Suitable values for R_1 and R_2 would be such that full scale deflexion is not exceeded when the power supply unit is delivering maximum output voltage (V_{max}).

Thus:

$$R_1 = R_2 = V_{max}/I_m$$

If the power pack is to have an output voltage range switch then extra switch sections could be provided with advantage to bring in lower values of R_1 and R_2 as V_{max} is reduced.

Earthing Switches and Links

It is considered undesirable to include switches on power supply units for the purpose of earthing an output terminal. Such switches, when inadvertently operated, have been known to earth one side of a supply which already had its other side grounded at some point in the external circuits. To say the least, unintentional short-circuits like this can be a source of inconvenience and embarrassment. The system suggested here shows clearly if one side is already earthed, and it allows the earth connexion to be made at the load end of the output leads.

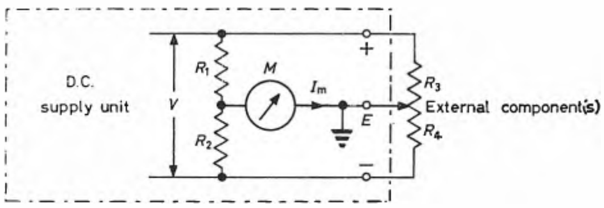


Fig. 4. Connexions for external resistance tests

Remote Units and Bench Supplies

Where supply units are situated at remote points from their respective loads the earth indicators will show the earth conditions at the other ends of the lines.

If one supply unit feeds a number of remote outlets, on various laboratory benches for example, duplicate output-earth indicator meters can be included with advantage at each bench position as shown in Fig. 3. Fuses and switches have been omittted from this diagram for clarification.

Additional Uses

Some additional uses have been found for power supply units fitted with this type of centre-zero voltmeter. If R_1 is accurately matched to R_2, and if M is sensitive, the power supply can be utilized for matching external resistors (R_3 and R_4 in Fig. 4) and for functionally checking potentiometer controls. Due regard must be paid, of course, to the wattage dissipated in the external components, and it must be seen that $V/(R_3+R_4)$ does not exceed the maximum output current rating of the power supply unit.

Substitution of a calibrated variable resistance unit for R_4, for instance, enables resistance measurements to be made when an unknown resistor (R_x) is connected in the position of R_3. In typical bridge fashion $R_x = R_3 = R_4$ when I_m is brought to zero by adjustment of R_4. Again attention must be given to wattage dissipations and the possibility of excessive current flow.

Qualitative insulation checks can be made by connecting the suspect leakage path between either output terminal and earth.

Acknowledgments

Credit is due to Mr. Keith Harvey, an AWRE apprentice, who constructed the prototype and to the Director of the establishment for granting permission for this article to be published.

An S.C.R. Pre-Regulator for a Transistorized Power Supply

By D. Ophir*

A reliable silicon controlled rectifier pre-regulator for a transistorized power supply is described. The pre-regulator is operated at high frequency from a direct current source.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

A REGULATED solid state power supply using a silicon controlled rectifier pre-regulator has been designed with specific emphasis on the overcoming of the usual limitations of such supplies: reliability of the s.c.r. control circuit and avoidance of highly disturbing noises caused by fast switching of high currents.

In conventional approaches to s.c.r. pre-regulators design, one of the following methods is used:

(1) Rectified line voltage is applied to the pre-regulator, while the main regulator controls the firing phase of the s.c.r.[1,2]. This method provides for a reliable turn-off characteristic of the series s.c.r., but requires very large tank capacitors at the input of the main series regulator (order of magnitude of 100 000μF).

(2) The second approach takes advantage of the fast switching ability of the s.c.r. by the pre-regulator being operated from a d.c. source, and switched at high frequency. The series pre-regulator uses an s.c.r. oscillator, which delivers during each cycle a certain constant amount of charge into a tank capacitor. The frequency of the oscillator is controlled by the main regulator. This configuration suffers from very high current peaks, and is not very reliable. For instance, the Morgan oscillator[3] which uses this approach suffers from these limitations. Reliability is particularly affected by spurious triggers which may appear before the turn off switching capacitor has recharged, and as a result the s.c.r. cannot be 'turned off' and remains continuously conducting.

The power supply described in this article combines high reliability and low noise characteristics, using a pre-regulator operated from a direct current source. The power supply can provide up to 5A at any voltage between 0 and 32V, while the pre-regulator maintains a voltage of approximately 2V across the series main regulator. Only a single series transistor mounted on a heat sink cooled by free air convection is used.

This article describes mainly the pre-regulator and its associated circuits. Most series regulators can be used with the same pre-regulator. The regulation obtained was better than $\pm 0{\cdot}1$ per cent for ± 10 per cent line variation and full load—no load switching. A peak-to-peak noise of less than 1mV was obtained. Better regulation can be achieved by the use of a higher class series regulator.

The S.C.R. Pre-regulator

The high reliability of the supply, as presented in the block diagram of Fig. 1, stems from the control of the s.c.r. pre-regulator conduction by independent continuous 'on' and 'off' orders. This is achieved by using two gated multivibrators which are operated through an 'OR' gate. The 'OR' gate is operated from a voltage discriminator connected across the main regulator.

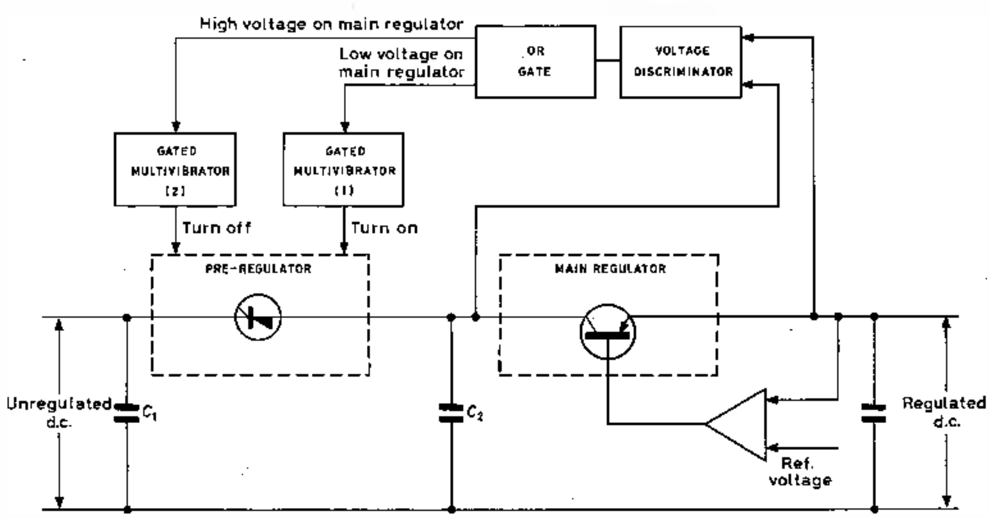
When the voltage across the main regulator is below a preset voltage (about 2V), the discriminator, through the 'OR' gate, enables the gated multi-vibrator No. 1 to run freely, thus providing a train of successive pulses which brings the pre-regulator into conduction. Charge is therefore transferred from C_1 to C_2.

Whenever the voltage across the regulator reaches a voltage higher than the preset voltage, the OR gate enables the gate multivibrator No. 2 to provide another train of pulses which stop the pre-regulator conduction. In this way the overall conduction time of the pre-regulator is controlled by the alternate orders of the two multivibrators. As the control consists of a train of successive commands, the circuit is independent of the failure of one command, or of the 'history' of the circuit before a command has been given.

S.C.R. Pre-regulator—Circuit Principle

The circuit principle of the pre-regulator is given in Fig. 2. To turn off the main s.c.r., a cathode pulse from a capacitive turn-off circuit is employed.

Fig. 1. Complete power supply



* *Israel Institute of Technology.*

When a turn-on command appears at the gate of SCR_1, the current starts to flow from C_1 into C_2. The rise time of this current and its maximum value are limited by the inductance L and the resistance R in series with SCR_1. At the moment that the voltage on the series regulator is above the preset value, a turn-off command is given. This command consists of a pair of pulses. One in each pair triggers SCR_2, and as a result SCR_1 is turned off. The coil L prevents the charge of C_3 from leaking via C_1 and C_2; that is, the whole charge of C_3 is available for the turn-off of SCR_1. The second pulse of each pair is applied between the base and the emitter of VT_1. This pulse is long enough to keep VT_1 non-conducting for a period which is longer than the discharge time of C_3. When C_3 cannot provide the necessary holding current for SCR_2, the latter turns off. SCR_1 remains non-conducting until a

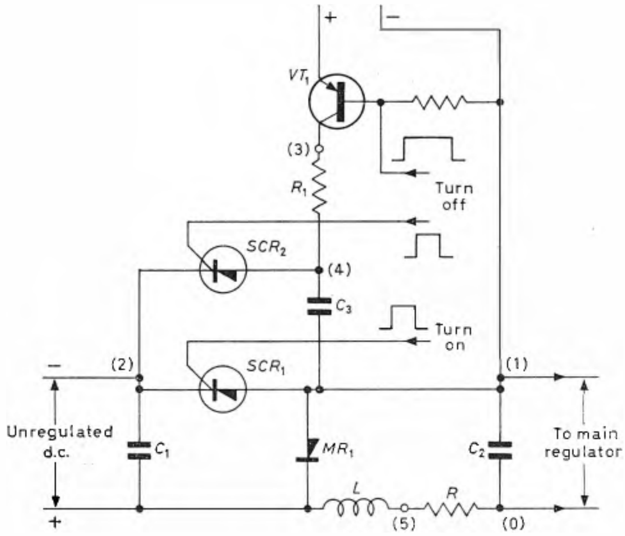


Fig. 2. S.C.R. pre-regulator circuit diagram—principle of operation

new 'on' order arrives. In the meantime VT_1 conducts again strongly and recharges C_3. The energy stored in the coil L during the conduction time of SCR_1 is returned into C_2 via diode MR_1. Waveforms at various points of the circuit are given in Fig. 3. For simplicity only one turn-off command is shown.

Mathematical Analysis of the Pre-regulator

The analysis is based on the equivalent circuit given in Fig. 4. The conditions, as stated below, for which the analysis was made were chosen in such a way as to provide the necessary information for 'worst case' checking.

(1) The output current is assumed to be switched from 0 to 5A (closing of S_2). If the output voltage is V, the load resistance can be given by $R_L = V/5$.

(2) It is assumed that the switch S_2 is closed at the moment that capacitor C_2 voltage, V_c, is at its minimum value, as preset by the voltage discriminator. The switch S_1 (which represents the series s.c.r.) will therefore be closed at the same time.

(3) The voltage on a series regulator should never decrease below approximately 1V. To avoid this situation the voltage discriminator is set to give a turn-on command whenever the voltage on the series regulator reaches 2V. This condition determines the value of the capacitor voltage at the time of the closing of the switches (zero time) as $V_{c(0)} = V +$ 2V. As the output voltage is constant, the series regulator can be given as a voltage source whose voltage is $(V_c - V)$. The unregulated d.c. voltage source is represented by the battery E.

The transformed loop equations are:

$$(SL+R+(1/SC))\,I_{1(mn)}-(1/SC)\,I_{2(mn)} = (E/s) - \frac{V+2}{S}$$

$$(-(1/SC))I_{1(mn)}+((1/SC)+(V/5))I_{2(mn)} = \frac{V+2}{S} -(V_{o(mn)}-(V/S))$$

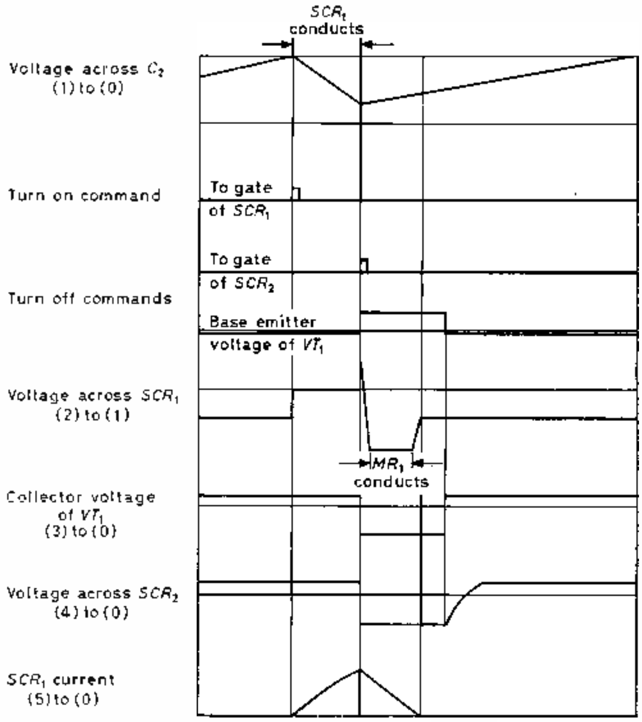


Fig. 3. Voltage waveforms in pre-regulator

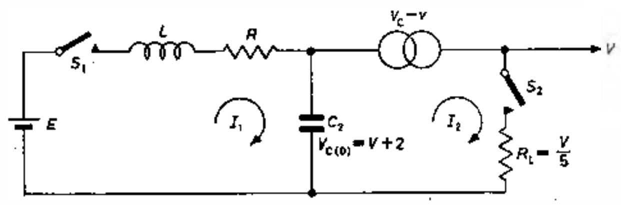


Fig. 4. Equivalent circuit

substituting (Condition 1):

$$I_{2(mn)} = 5/S,$$

it is possible to show that

$$V_{o(mn)} = (V+2)/S - 5/(S^2C) + \frac{E-V-2}{LCS[(S+(R/2L))^2+1/LC-R^2/4L^2]} + \frac{5}{S^2LC^2[(S+(R/2C))^2+(1/LC-R^2/4L^2)]}$$

The series regulator voltage is given by:

$$V_c = E-5R+(5R+V+2-E)\exp(-(R/2L)t) \cos\sqrt{(1/LC-R^2/4L^2)}t +$$

$$[(5/2)(R^2/L)-5/C+R/2L(V+2-E)]\exp(-(R/2L)t)\,.\,\frac{\sin\sqrt{(1/LC-R^2/4L^2)}t}{\sqrt{(1/LC-R^2/4L^2)}}$$

The s.c.r. current is given by

$$I_1=\frac{E-V-2}{L}\left[\exp(-(R/2L)t)\,.\,\frac{\sin\sqrt{(1/LC-R^2/4L^2)}t}{\sqrt{(1/LC-R^2/4L^2)}}\right]-5\left[\exp(-(R/2L)t)\cos\sqrt{(1/LC-R^2/4L^2)}t+\frac{(R/2L)\exp(-(R/2L)t)\sin\sqrt{(1/LC-R^2/4L^2)}t}{\sqrt{(1/LC-R^2/4L^2)}}-1\right]$$

The values that are used in the final power supply are:

$L = 2\text{mH}$
$C = 4\,000\mu\text{F}$
$R = 1\Omega$
$E = 50\text{V}$ (for 220V mains voltage).

Two curves which are of interest for the worst case design are given:

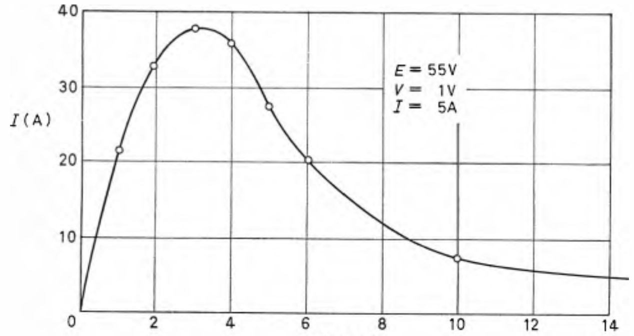


Fig. 5. S.C.R. current waveform

(1) *The S.C.R. Current*

This current has its shortest rise time and can reach its maximum value in the following conditions:

The output voltage is low (V=1V). The output current is switched from 0 to 5A, and the mains voltage increased by 10 per cent (E=55V). For these conditions the current is given by:

$$I = 104\,e^{-250t}\sin 250t+5-5\,e^{-250t}\cos 250t.$$

This waveform is given in Fig. 5. The current will of course never reach the maximum value of 38A, and the turn-off command will cut the current rise after a short period. It is therefore better to state that the current rises at 22A/msec.

The maximum current of 38A is considered only as a precaution against overload.

(2) *The Series Regulator Voltage*

The curve was drawn for the following conditions:

The output voltage is at maximum V=33V. The current is switched from 0 to 5A under the same conditions as described under (2) above. For these conditions the worst case will be when the mains voltage decreases by 10 per cent (E=45V).

The pre-regulator voltage is given by

$$V_o = 40 - 5\,e^{-250t}\cos 250t - 10\,e^{-250t}\sin 250t.$$

This waveform is given in Fig. 6. It can be seen that the voltage on the series transistor will never decrease below 1·2V, even under worst case conditions.

Complete Power Supply Circuits

The complete pre-regulator circuit is shown in Fig. 7. Transistors VT_2, VT_3, VT_4 fulfil the function of the voltage discriminator. VT_5 and VT_6 form the turn-on command multivibrator, and the turn-off multivibrator includes transistors VT_7 and VT_8. Diodes MR_2 and MR_3 form the OR gate. Transistors VT_9, VT_{10} and VT_{11} operate as a single shot multivibrator. Whenever this single shot multivibrator is activated, no holding current is available for SCR_2, and SCR_2 is turned off after C_3 has discharged. MR_4 and MR_5 form an additional gate for the turn-off multivibrator. They will enable this multivibrator to deliver the train of

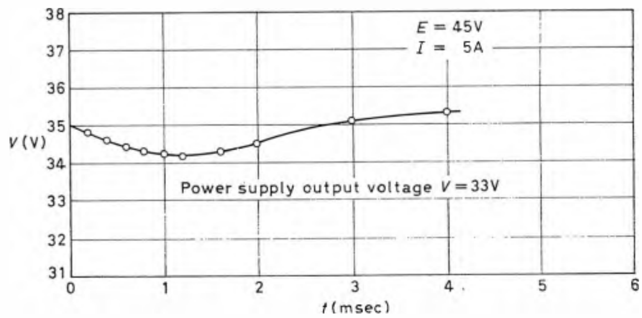


Fig. 6. Main regulator input voltage

turn-off commands only when SCR_1 is conducting. If the gate were not incorporated, C_3 would be discharged at the rate fixed by the turn-off multivibrator. For low currents and voltage output this charge can be high enough to charge C_2 above the desired voltage. In this way only one successful turn-off pulse can be given in each cycle.

The main limitation for a high rate of operation of the control circuit is set by the time required to recharge C_3. Consequently, one cannot take full advantage of the high regulation potentiality provided by a low hysteresis voltage discriminator because the rate of 'on' and 'off' cycles will be so high that C_3 will not have the necessary time to recharge to the desired value.

A compromise is obtained by the use of a sensitive discriminator, but its output is delayed by an RC circuit (R_4C_4) before it reaches the gated multivibrator. This lowers the overall regulation, but assures the reliable operation required.

In Fig. 8 the main regulator is shown. When the output voltage is high and the power supply is short-circuited, a momentarily high voltage, equal to the voltage on C_2 before the short-circuit occurred, will appear across the series transistor. A Zener diode MR_6 is connected from the collector of the series transistor into the regulator amplifier. This diode avoids extensive power dissipation in the series transistor, which would appear if C_2 could discharge under short-circuit conditions via the series transistor. This arrangement will first reduce the output voltage to zero without output current, then C_2 will discharge through its bleeder. After the series transistor reaches the preset voltage region (about 2V), the output current increases to the preset short-circuit current (given

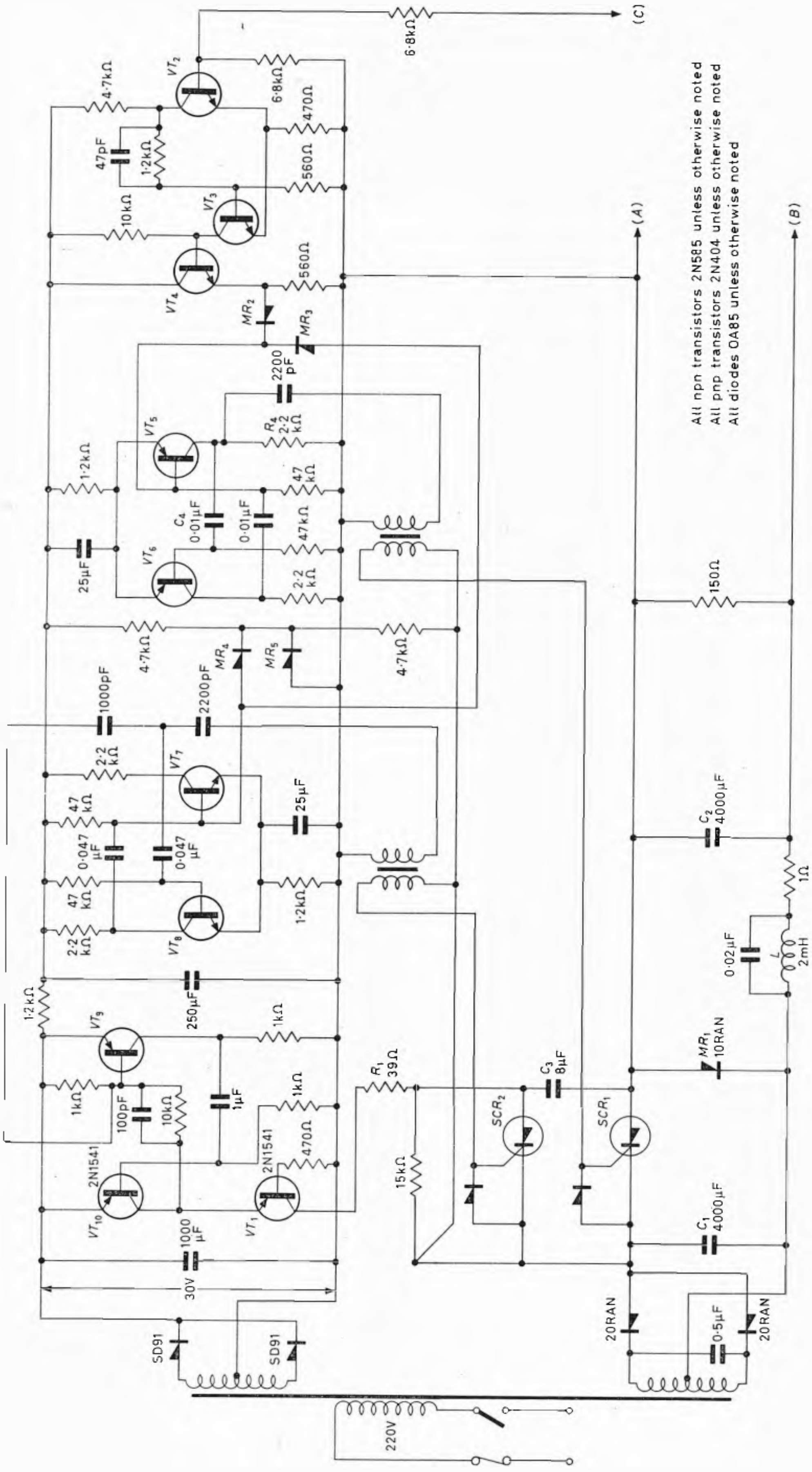


Fig. 7. Pre-regulator

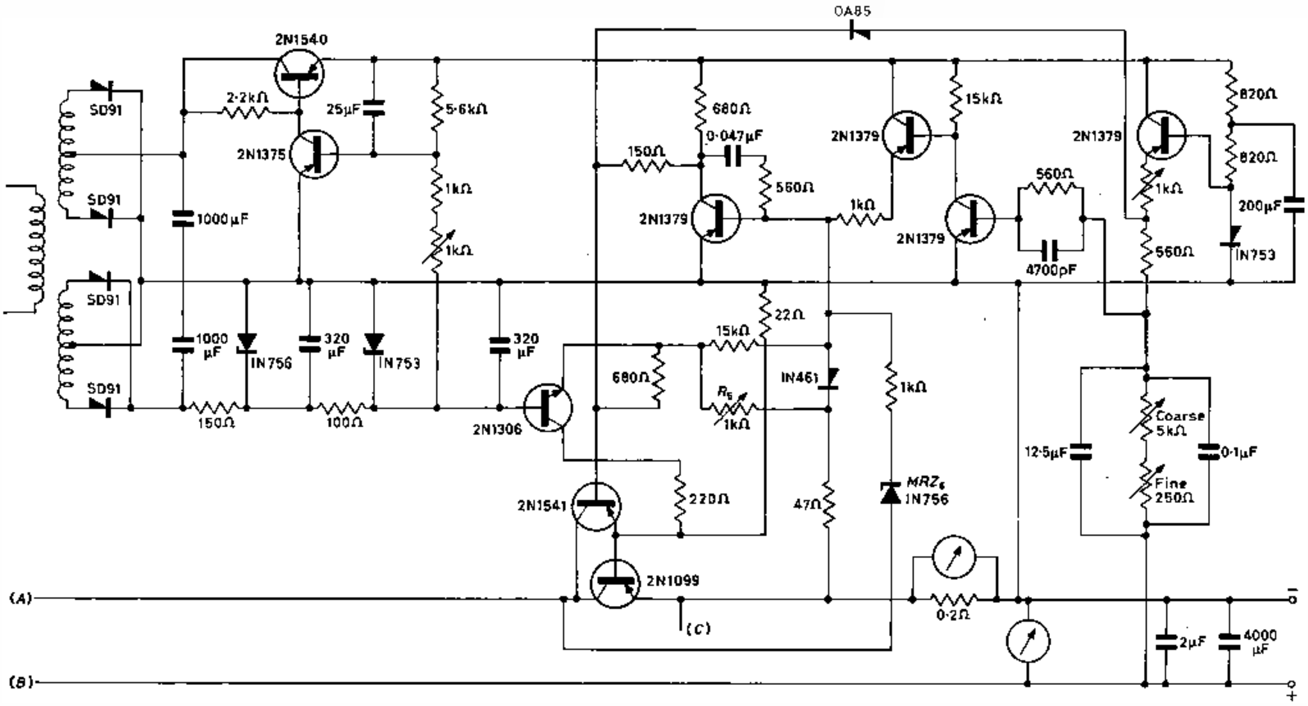


Fig. 8. Main regulator

by R_5). It is therefore enough to select a series transistor that is able to withstand high voltage conditions without at the same time admitting high currents. After the short-circuit is removed, the power supply recovers immediately. The current waveform of SCR_1 is given in Fig. 5. Under these conditions the peak current of 38A can be reached.

The stability obtained is better than $\pm 0{\cdot}1$ per cent. The noise appears as pulses of less than 1mV peak-to-peak. The mains frequency content in the noise is less than 100μV peak-to-peak.

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A 'Door-Bell' for the Deaf-Blind

To those who are both deaf and blind—and especially to those who live alone—one of the most worrying features of day-to-day life is the inability to know when there is a caller at the front door.

Many attempts have been made to solve the problem by a variety of means, such as electric fans to create air movement, scent sprays to provide aromatic indication and devices to vibrate windows, furniture or floors. All have proved unsuccessful. Now, The Royal National Institute for the Blind has found a solution.

The first of its kind in the world, the system is simple, inexpensive and entirely acceptable to the deaf-blind, who have been closely concerned with every stage of its development. Mr. A. R. Sculthorpe, M.B.E., Secretary of the National Deaf-Blind Helpers' League, who is himself totally deaf and blind and in whose home a pilot installation was made, considers it to be "An unqualified success."

The system is based on an idea put forward by the G.P.O. Research Branch, Dollis Hill, where a demonstration of its possibilities was given some 18 months ago, and which has now been developed by the R.N.I.B. into a practical, working proposition. The first prototype was produced six months after this demonstration and has, in principle, remained unaltered.

The equipment consists of two component parts, the first static, the second mobile. The former comprises a series of loops of wire or copper strip unobtrusively encircling the living accommodation, connected to a mains operated transmitter which, when actuated by the bell-push, feeds current to the loops, thereby creating an electromagnetic field throughout the home. The latter, which is worn by the user, consists of a pocket-size transistorized hearing-aid receiver, sensitive to this field, which operates not an ear-piece, but a small vibrator worn on the finger, experiments having shown this area of the body to be most suitable from the viewpoints of tactile response and convenience.

Tests to determine the optimum frequency had shown that, for tactile response, this lay between 200c/s and 300c/s and that maximum sensitivity of the receiver occurred at the higher end of this range. Low ultimate cost, however, had, throughout the entire development of the system, been of prime importance. The need for a cheap transmitter, therefore, lead to the preparation of a specification for a saturable core transformer locked to a multiple of the mains frequency. As the limit in this direction appeared to be 250c/s, and as this fell within the acceptable tactile range, this frequency was adopted as the standard.

The vibrator raised special problems, as the operative power delivered by the hearing aid was so low. This problem was finally overcome by modifying the spring assembly within a bone conductor unit to create natural oscillation at 250c/s.

Finally, the size and shape of the loop circuit, together with the type of cable necessary, were established and, to accommodate the wide range of residential layouts which would be involved, a compromise specification was agreed.

The Ministry of Health has now agreed to supply sufficient receivers to permit a wide-scale field trial, to a maximum of 200 installations, to be carried out early in 1965 and has invited a number of local authorities in England and Wales to participate in this. Local welfare authorities taking part will notify to the R.N.I.B. details of deaf-blind persons to be provided with installations and after a period of six months a questionnaire will be sent to these experimental users. Information received will be used for evaluation purposes to ensure that the device meets all operational requirements before it is made nationally available.

A Reversible Decade Counter with D.C. Logic Reset

By J. H. Smith*, A.M.I.E.R.E., M.S.I.T.

A reliable reversible decade counter with d.c. gated forced reset, is described. The counter is arranged to give the same output code in either counting direction. The development of a similar decade is also described where the decimal output for both counting directions is in the 8421 binary code. These counters are not self-complementing.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

A NUMBER of reversible decade counters has been described[2,3,4] but are unsuitable for the application required. The counters to be described are designed for use in laboratory equipment where the basic facilities are:

(1) A reversible counter comprising two decades, two binary and one divide by five.

(2) Automatic reversing is used, controlled by a bistable circuit.

(3) All outputs of the counter are decoded by a single decoder, e.g. four (4) is always represented by 0100 whatever the counting direction.

Examination of the literature showed that the counters described by Richards[2] were unnecessarily complex for a practical design. Carlsons[3] counter was complicated and the reset arrangements proved unreliable in operation. The counter would be suitable, with further development, if a complementary output was necessary. The counter described by Scollar[4] was unsuitable due to the decoding arrangements required by the laboratory equipment.

Reversing Binary

The basic reversing binary counter (Fig. 1) has been described and used extensively[1,2,3,4].

Any binary stage may be used in the system described providing it has 'a.c.' triggering with 'd.c.' set and reset inputs. The system described uses Combi-Elements manufactured by Mullard Equipment Ltd.

The binary stages in a divider are designated $A.B.C.D.$, the two outputs being A, $\overline{A}$, B, $\overline{B}$, etc. The outputs A, B, C, D, will represent (1) when the counter is set, and (0) when it is reset. The set and reset circuits are driven from a (1) and are designated set A. or reset $.A$.

The gating arrangement shown in Fig. 1 provides a positive transient from either A or $\overline{A}$ depending upon the position of the forward reverse switch. Referring to A, the trigger signal to the following binary may be regarded as the (1) ⟶ (0) transient of A when counting forward and as the (0) ⟶ (1) transient of A when counting in the reverse direction. The reverse transient is, of course, actually the (1) ⟶ (0) transient of the $\overline{A}$ output. If a table of outputs from a series of binary stages A, B, C, D, is drawn, it is clear that the counter is reversible, Table 1.

The counter circuit shown in Fig. 1 is capable of counting up to 100kc/s but the reversing gates must be switched at least 60μsec before the following signal. This ensures that the gating circuits have settled into the new mode of operation. It is found that for maximum reliability the reversing signals must be 0V and $-6V \pm 0{\cdot}5V$ from a low impedance source. The final instrument uses a flip-flop reversing switch driving a complementary emitter-follower.

Reversing Decade

Four reversible binary stages can be made into a decade by using forced reset. In the following arguments all signals are logically related to the A, B, C, D, outputs of the binary stages. Therefore, the stages may be considered as triggering on the (1) ⟶ (0) transient in the forward direction and on the (0) ⟶ (1) transient when reversed. If

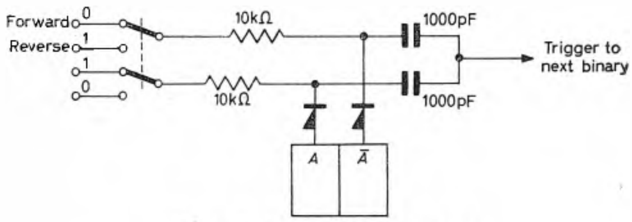


Fig. 1. Basic reversible binary stage

careful use is made of these triggering transients it is possible to set the counters forward or backwards at will. An example of the technique is shown in Table 2.

Table 2 shows how state 9 sets the counter forward to state 15, by setting B and C into the (1) condition. State 14 is used to set back to state 8 by resetting B and C into the (0) condition. These two operations have no effect on the condition of $.D.$ and A.

It is possible to set forward for a decade in four positions these are, 0 ⟶ 6, 1 ⟶ 7, 8 ⟶ 14 and 9 ⟶ 15: while the reverse direction has only two 14 ⟶ 8 and 6 ⟶ 0. The arrangement shown in Table 2 is used in the laboratory apparatus.

Reset Gating

From Table 2, it can be seen that the reset transient in the forward direction is from 9(1.0.0.1.). This signal may be formed by a gate $A.\overline{B}.\overline{C}.D.$ and the output of this gate could be used to set $B.$ and $.C.$ into the (1) condition. If this signal were used to set B and C into the (1)

TABLE 1
Binary Counter Output
(forward and reverse trigger)

	OUTPUT					OUTPUT			
Forward	*D.*	*C.*	*B.*	*A.*	*Reverse*	*D.*	*C.*	*B.*	*A.*
(1) ⟶ (0)	0	0	0	0	(0) ⟶ (1)	1	1	1	1
Trigger	0	0	0	1	Trigger	1	1	1	0
	0	0	1	0		1	1	0	1
	0	0	1	1		1	1	0	0
	0	1	0	0		1	0	1	1
	etc.					etc.			

* *Gillette Safety Razor Co.*

condition, it would vanish as soon as B or C were set. Such a reset arrangement would be unreliable as the counter could set into conditions 1011, 1101 or 1111, "11, 13 or 15". The reset gate must be arranged to give an output on 9, 11 and 13 to avoid resetting into an incorrect condition.

The reset gate is given by: $A\,\bar{B}\,\bar{C}\,D + A\,\bar{B}\,C\,D + A\,B\,\bar{C}\,D$

which simplified $= A\,D\,(\bar{B}\,\bar{C} + \bar{B}\,C + B\,\bar{C})$

$A\,D\,[\bar{B}\,(\bar{C} + C) + B\,\bar{C}]$

$\bar{C} + C = 1$

$\therefore$ Reset Gate $= A\,D\,(\bar{B} + B\,\bar{C})$

The bracketed expression $(\bar{B} + B\,\bar{C})$ will be a (1) if either $\bar{B}$ is (1) or $\bar{C}$ is (1) because if B is (0) then $\bar{B}$ is (1); if $\bar{B}$ is (1) then $\bar{C}$ determines the output. The expression is identical to $(\bar{B} + \bar{C})$. Therefore the reset gate simplifies to:

$$A.D.\,(\bar{B} + \bar{C}) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

This expression fully describes the reset gate which sets B and C into the (1) condition.

Similar reasoning to the above arguments show that the reset gate for reverse counting must gate "14, 12, and 10" and is described by the expression:

$$\bar{A}.D.\,(B + C) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

The validity of these expressions can be established by referring to Table 2.

Reset Circuit

The forward reset, expression (1), may be formed by the circuit shown in Fig. 2.

The practical circuit arrangement for forward and reverse reset is shown in Fig. 3.

Two identical circuits are required, one has an input $A.D.\,(\bar{B}+\bar{C})$ and sets B and C in the (1) condition, forward reset. The other has inputs $\bar{A}.D.\,(B + C)$ and resets B and C into the (0) condition, reverse reset.

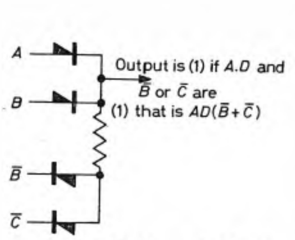


Fig. 2. Basic reset circuit

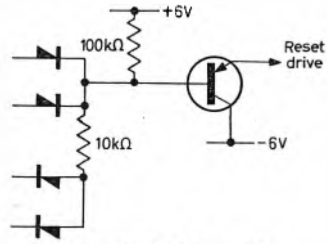


Fig. 3. Practical reset gate

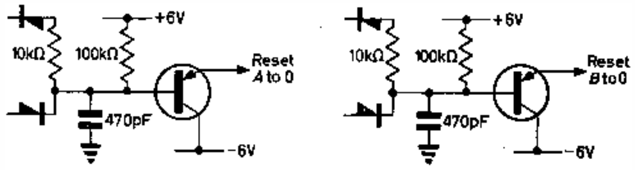


Fig. 4. Divide by five reset arrangement

This method of reset gating is infallible provided that the counters to be set or reset are those which appear in the OR bracket, so the reset signal can always be present when it is required. Therefore a necessary theorem for d.c. gated reset is that only the bracketed OR stages can be reset. When these conditions are met a fast reliable reset pulse will be generated, the reset time being determined only by circuit design of the binary stage.

TABLE 2
A Binary Decimal Counter Unit

		D.	C.	B.	A.	
0		0	0	0	0	
2		0	0	0	1	
2		0	0	1	0	
3		0	0	1	1	
4	Reverse Count	0	1	0	0	Scale Forward
5	trigger (0) ⟶ (1)	0	1	0	1	trigger (1) ⟶ (0)
6		0	1	1	0	
7		0	1	1	1	
8		1	0	0	0	
9		1	0	0	1	Transient State
10		1	0	1	0	
11		1	0	1	1	
12		1	1	0	0	
13		1	1	0	1	
14	Transient State	1	1	1	0	
15		1	1	1	1	

The Divide by Five

The laboratory equipment for which these counters were designed required a divide by five which gives zero to four in pure binary code. The arrangement required is shown in Table 3.

Examination of the reverse direction shows that A and B must be reset to (0) and the signal must be derived from 7, 6 and 5 thus:

$$A\,B\,C + \bar{A}\,B\,C + A\,\bar{B}\,C$$
$$= C\,(A\,B + \bar{A}\,B + A\,\bar{B})$$
$$= C\,(B\,(A + \bar{A}) + A\,\bar{B})$$
$$= C\,(B + \bar{B}\,A)$$

The bracketed expression simplifies as shown for expression (1). Therefore the reset gate is:

$$C\,(B + A) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Expression (3) is used for resetting $A\,B$ to zero on the reverse count.

A similar argument may be applied to the forward reset, thus 5, 6 or 1 must reset $A.B.C.$ to zero.

TABLE 3
A Divide by Five

		C.	B.	A.	
0	Reverse	0	0	0	Forward
1	Trigger (0) ⟶ (1)	0	0	1	Trigger (1) ⟶ (0)
2		0	1	0	
3		0	1	1	
4		1	0	0	
5		1	0	1	Transient State modified forward
6		1	1	0	
7	Transient State	1	1	1	reset arrangement
0		0	0	0	

The reset gate is given by $A\,\bar{B}\,C + \bar{A}\,B\,C + A\,B\,C$ which is identical to the previous expression, hence $C(B + A)$.

This reset arrangement does not satisfy the general theorem that only binaries in the bracket can be reset and the gate will in fact lock on to 4 (1.0.0.) when operated in the forward direction. There is no simple logical arrangement which will enable a reliable forward and reverse divide by five to be made. However, use can be made of the fact that the action of setting a flip-flop from (1) ⟶ (0) in the forward direction triggers the subsequent stage. The following sequence of events is used.

(1) Form gate *A.C.* (which occurs first on No. 5).

(2) Reset *A* to (0).
as *A* changes from (1) it triggers *B* into state (1) and the counter is set into state 6, (1, 1, 0.)

(3) Form gate *B.C.*

(4) Reset *B* to (0).
as *B* changes from (1) ⟶ (0) it triggers *C* into the (0) condition leaving the counter in the required state 0 0 0. The sequence is shown on Table 3 by dotted arrows; labelled, modified forward reset.

This arrangement, as described, is not infallible due to the fact that the binaries are triggered into state (1) and at the same time gated to form the reset signal. The two signals can cancel out and the system will again lock into state 4. This is overcome by delaying the reset signal as shown in Fig. 4, where a capacitor across the gate acts as the delay element. A more economic arrangement would be to use a slow speed reset transistor.

The reverse reset arrangements are unaffected by these circuit modifications and are substantially the same as expression (3). Therefore the system is capable of giving reliable results in the forward and reverse directions.

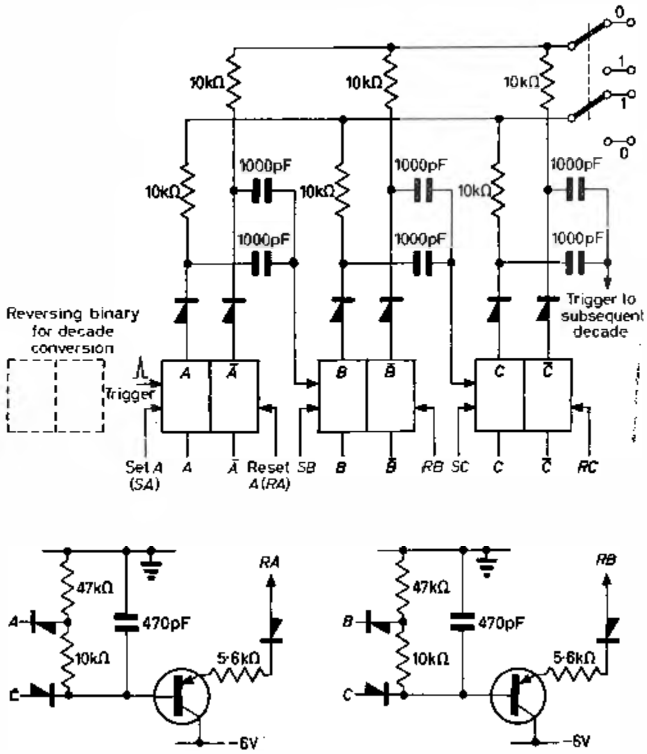


Fig. 6. Divide by five

Reversing Decade 2

It is of interest to note that this divide-by-five may be converted to a decade by the addition of a further binary. This decade will have a few advantages over decade 1, these are:

(1) Four diodes used for reset as opposed to eight.

(2) Slow speed transistors may be used in the reset circuits.

(3) The output is coded in pure binary 0-9.

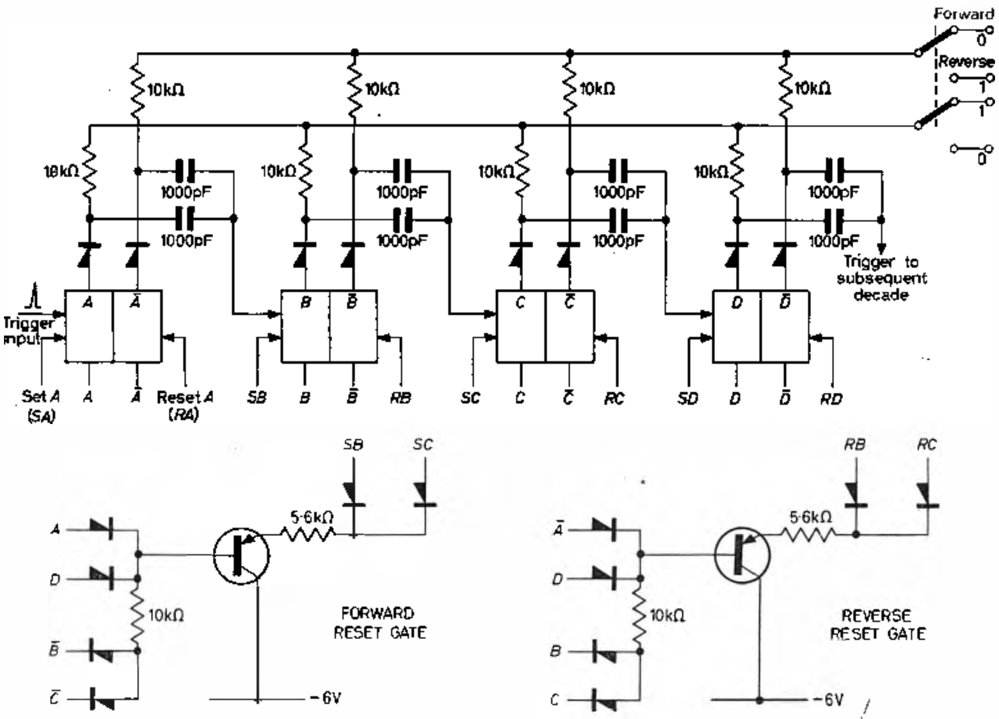
The circuit has the disadvantage of being slower to reset and will take twice as long as decade .1. due to the two stages being reset in sequence.

A full circuit of decade .1. and the divide-by-five is shown in Figs. 5 and 6.

Conclusions

A technique has been described for the design of binary/decade reset circuits. The decade circuits shown have been found useful and reliable in many applications. It has also been shown that the system can be extended to a scale of five, although modifications are necessary. The basic technique is not limited to reversible decades but may be employed for counters of different radix, and may also be used extensively with non-reversing counters.

Fig. 5. Reversing decade



Acknowledgments

The author wishes to thank Mr. P. S. Williams and the Directors of Gillette Industries Ltd for permission to publish this article.

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Operation of a Magnetically Coupled Multivibrator with Square BH Loop Cores

By G. Smiljanic*

Some features of a frequency controlled magnetically coupled multivibrator are considered for the case of driving transformer cores having square hysteresis loops. It is shown that the frequency of the multivibrator depends linearly on the input voltage. The frequency can be adjusted by changing resistors, stabilized by Zener diodes against variations of the input voltage or temperature, and varied in a wide range by a d.c. control current, keeping the square waveshape of the output pulses at constant amplitude.

(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 72)

In an earlier article a frequency controlled magnetically coupled multivibrator was proposed and the principle of its operation was explained[1]. In the present article certain results obtained by the proposed magnetically coupled multivibrator are given for the case when the transistor base-driving transformers have rectangular hysteresis loop cores.

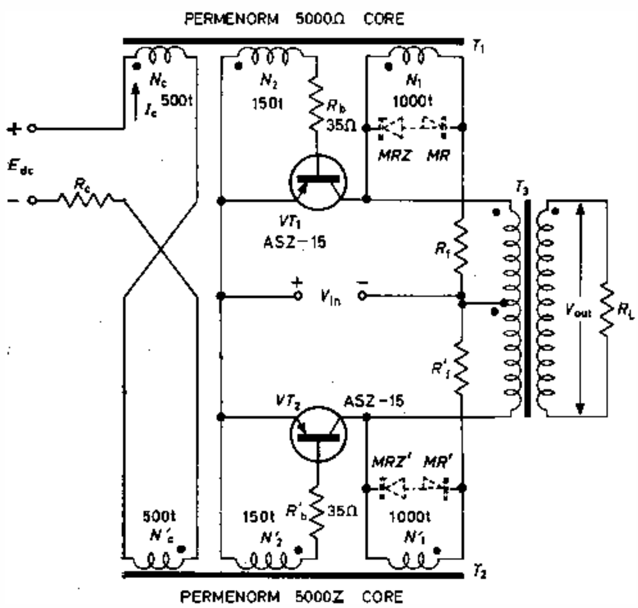

Fig. 1. The magnetically coupled multivibrator

The frequency controlled magnetically coupled multivibrator is shown in Fig. 1. Neglecting for a moment the control windings N_c and N_c' the operation is as follows. The transistors VT_1 and VT_2 conduct successively. While VT_1 is conducting, VT_2 is cut off by the voltage across the winding N_2', the core of the transformer T_1 (Fig. 1) is 'excited' being driven from $-B_m$ to $+B_m$ (Fig. 2), and the core of the transformer T_2 is 'reset' being driven from $+B_m$ to $-B_m$. When saturation of the transformer core T_1 is reached (T_1 determines the duration of this period), the transistor VT_1 cuts off and VT_2 begins to conduct. The next period duration is determined by the transformer T_2. Thus the multivibrator, which is symmetrical, oscillates at the 'base' frequency, determined by the saturation-to-saturation magnetic flux change in T_1 and T_2. If a d.c. control current is now applied to the windings N_c and N_c', it always aids to drive the core which determines the pulse duration. By changing the control current, the amount of the magnetic flux taking part in the operation can be changed, and the frequency varied. Successive conductions and cut-offs of the transistors VT_1 and VT_2 result in a square waveshape of the output pulses, transmitted to the load by the linearly operating transformer T_3.

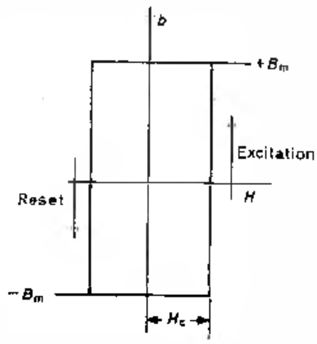

Fig. 2. The BH loop

Operation of the Multivibrator with Rectangular BH Loop Cores

The multivibrator has the most useful features and is therefore most frequently used in the case when the transformer cores T_1 and T_2 are made of square *BH* loop material (Fig. 2). In that case the total magnetic flux, which determines the conduction period duration, is sharply limited by the amount $2\Phi_m$, if the control windings N_c and N_c' are omitted or the control current is zero. Because of the constant coercive force (Fig. 2), the magnetizing current flowing through the resistors R_f or R_f' is constant during the conduction period. Further, if the resistors R_b and R_b' are sufficiently large, so that the nonlinearities of the emitter-base resistances can be neglected, and the current, flowing through the resistors R_f and R_f', necessary to compensate the influence of eddy currents in the cores is negligible, then the voltage E across the windings N_1 or N_1', during the conduction period, is also constant. Thus, assuming that the excited core is in the state $-B_m$ when the conduction period begins, it can be written:

$$\int_0^T E\,.\,dt = N_1 \int_{-\Phi_m}^{+\Phi_m} d\phi \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Solving equation (1), one obtains expressions for the

* *Institute 'Ruder Boskovic', Zagreb, Yugoslavia*

duration of the conduction period T and the frequency f

$$T = \frac{2N_1\Phi_m}{E} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

$$f = \frac{E}{4N_1\Phi_m} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Assuming that the emitter-collector voltage of the conducting transistor is negligible in comparison with V_{in}, because the transistor, being able to handle any necessary

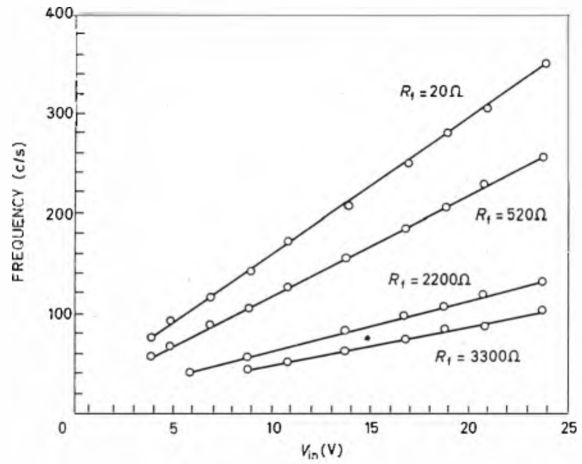

Fig. 3. Frequency dependence on voltage

current, always operates in the saturation region, the voltage E is:

$$E = V_{in} - (H_c l/N_1) \,.\, R_f - I_b \,.\, R_f \quad \ldots\ldots\ldots \quad (4)$$

where I_b is the current which flows during the conduction

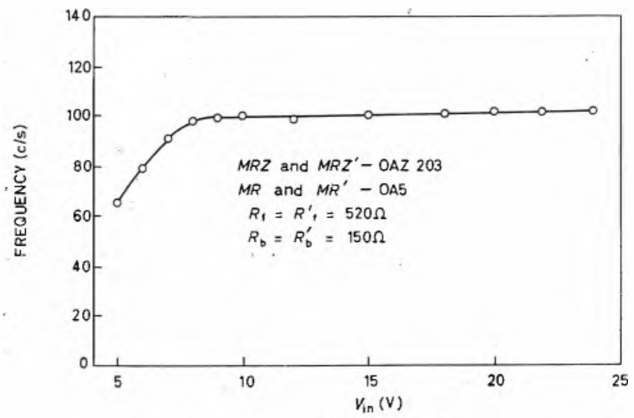

Fig. 4. Frequency dependence using Zener diode

period through the resistor R_f or R_f' because of the transistor base current, and l is the mean length of magnetic flux lines of the transformer T_1 or T_2.

Finding I_b and substituting it in equation (4) one obtains

$$E = \frac{V_{in} - R_f \,.\, (H_c l/N_1)}{1 + (R_f/R_b) \,.\, (N_2/N_1)^2} \quad \ldots\ldots\ldots\ldots \quad (5)$$

From equations (3) and (5) it follows:

$$f = \frac{V_{in} - R_f \,.\, (H_c l/N_1)}{4N_1\Phi_m\,[1 + (R_f/R_b) \,.\, (N_2/N_1)^2]} \quad \ldots\ldots \quad (6)$$

or

$$f = k_1 V_{in} - k_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

where k_1 and k_2 are constants.

Equations (6) and (7) show that the frequency depends linearly on V_{in}.

The measured frequency dependence on V_{in} of the multivibrator from Fig. 1 (which satisfies the above assumptions) for different values of the resistors R_f and R_f' is shown in Fig. 3. The frequency depends linearly on V_{in}, as it follows from equations (6) and (7), and as it was the case with the multivibrator proposed by Royer[2] and Uchrin and Taylor[3]. However, in the case of the multivibrator shown in Fig. 1, at constant V_{in}, the frequency can be adjusted in a wide range by change of the resistors R_f and R_f'. The efficiency of the multivibrator is not reduced materially when resistors of larger values are used,

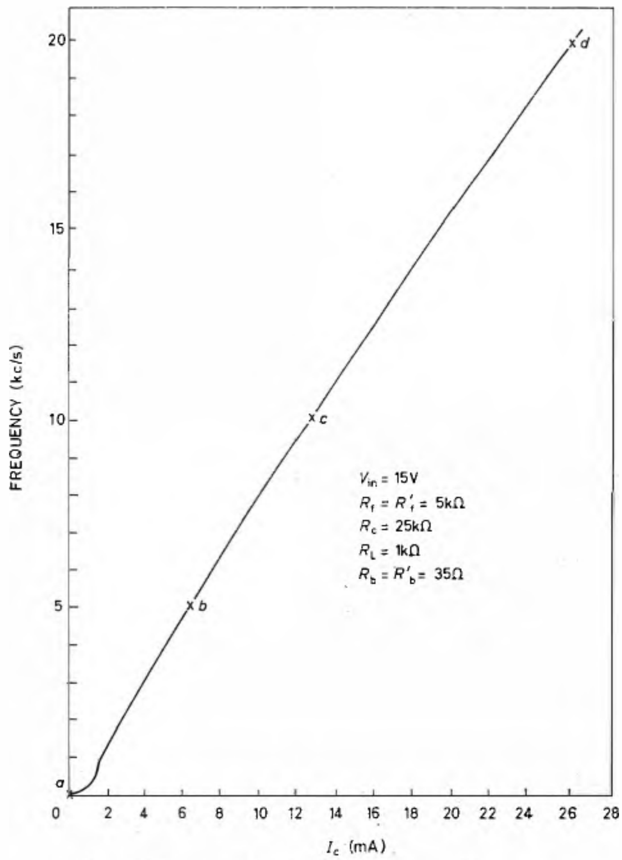

Fig. 5. Frequency dependence on d.c. control current

because only a small current flows through them, if the BH loops of T_1 and T_2 are not only square but also narrow, as is usual.

The voltage E across the windings N_1 and N_1' can be stabilized during excitation period by Zener diodes MRZ and MRZ' as shown in Fig. 1 by dashed lines. The diodes MR and MR' in series with the Zener diodes are necessary to avoid short-circuiting of the windings N_1 and N_1' during the reset periods of the transformer cores. To the constant voltage across the windings N_1 and N_1' corresponds a constant period duration and frequency as shown by equations (2) and (3). Thus, the period duration and frequency dependence on the input voltage are almost completely eliminated when Zener diodes are used to stabilize the voltages across N_1 and N_1'.

Using the Zener diode OAZ203 in series with an OA5, the frequency dependence of the multivibrator in Fig. 1 on V_{in} is shown in Fig. 4. (The control windings N_c and N_c' are, as in the previously discussed case, un-

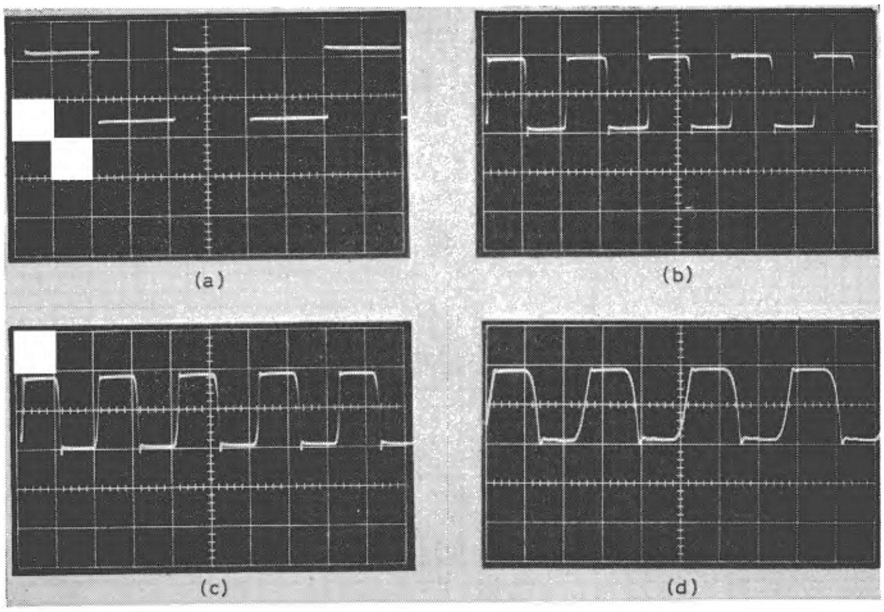

Fig. 6. Output pulses at points shown in Fig. 5
Horizontal scale: (a) 5msec/div, (b) 0·1msec/div, (c) 50μsec/div, (d) 20μsec/div; Vertical scale: 20V/div

necessary). To stabilize the frequency, the Zener diode has to stabilize the voltage across N_1 and N_1' during the whole conduction period. This is possible only for input voltages above some minimum value of V_{in}, which is about 9V in the present case. The frequency change in Fig. 4 amounts to 4 per cent for an input voltage change from 9 to 25V.

Although no measurements of frequency dependence on temperature have been made, it can easily be seen that temperature variations of the total magnetic flux $2\Phi_m$ can be compensated for by corresponding temperature changes of the voltage E, if suitable temperature coefficients of the Zener diode and the diode in series with it are chosen to match the temperature coefficient of the magnetic flux. In such a way a very stable frequency can be obtained when V_{in} is constant.

So far the control windings N_c and N_c' have been omitted as unnecessary, and it has been assumed that the frequency of the multivibrator is the base frequency determined by the magnetic flux change from saturation to saturation. But the control windings N_c and N_c' give the multivibrator the possibility of frequency change by means of a control current, changing the amount of the magnetic flux which determines the frequency.

The frequency dependence of the multivibrator, from Fig. 1, on the d.c. control current is shown in Fig. 5. The base frequency ($I_c = 0$) is about 50c/s. At a nearly maximum control current ($I_c = 26$mA) the frequency is about 20kc/s, i.e. a frequency change of 1:400 is obtained by the use of appropriate transformer cores, and in a wide range the output pulses are almost ideally square with constant amplitude. The output pulses, taken at points marked by *a*, *b*, *c* and *d* in Fig. 5, are shown in Fig. 6(a), (b), (c) and (d), respectively. In Figs. 6(a), (b) and (c) the pulses are rectangular, so that the wave shape of the pulses in Figs. 6(b) and 6(c), obtained at 100 and 200 times base frequency respectively, differ very slightly. Thus, it can be concluded that the change of frequency by the d.c. control current does not impair the waveshape of the output pulses, except that the rise and fall times, which remain approximately constant in the whole range, amount to a larger percentage of the total pulse duration at shorter pulses, as can be seen from Figs. 6(b), (c) and (d).

Moreover, Fig. 6(d) shows that the frequency of 20kc/s could be raised still more by a larger control current, reducing still further the flat top of the output pulses. But in that case the output pulse ceases to be rectangular.

Conclusions

It is shown that the frequency controlled magnetically coupled multivibrator, built with driving transformers having rectangultar *BH* loops, has the following features:

(1) The frequency depends linearly on the input voltage. At constant input voltage the frequency can be set in a wide range by appropriate change of the resistors.

(2) The frequency can be made independent of the input voltage changes by stabilizing the driving transformers using Zener diodes.

(3) By the use of Zener diodes the frequency can be stabilized against temperature variations.

(4) At constant input voltage and thus constant output pulse amplitude, the frequency can be varied in a wide range by a d.c. control current. The output waveform obtained is an almost ideally rectangular pulse.

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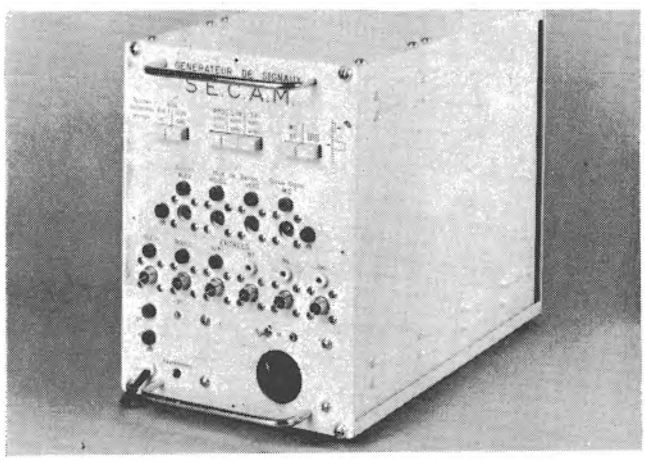
2. ROYER, G. H. A Switching Transistor D.C. to A.C. Converter having an Output Frequency Proportional to the D.C. Input Voltage. *Trans. Amer. Inst. Elect. Engrs. (Commun. and Electr.).* 74, 322 (1955).
3. UCHRIN, G. C., TAYLOR, W. O. A New Self-Excited Square-Wave Transistor Power Oscillator. *Proc. Inst. Radio Engrs.* 43, 99 (1955).
4. VAN ALLEN, R. L. A Variable Frequency Magnetic-Coupled Multivibrator. *Trans. Amer. Inst. Elect. Engrs. (Commun. and Electr.).* 74, 356 (1955)

A Colour Television Signal Generator

A fully transistorized generator for the production of signals necessary for setting up a SECAM colour television receiver has been produced by Compagnie Française de Television of Levallois-Perret, Seine, France.

The generator, type GS-10, is intended for use in the laboratory stages of design and prototype production of colour television receivers working on the SECAM system and together with an appropriate synchronizing pulse generator it produces the following signals: (1) A black-and-white signal: convergence pattern. (2) The three bar test pattern primary signals. (3) A SECAM bar pattern signal. (4) A SECAM coded signal of external primary signals. (5) A series of typical signals, selected by push-buttons, for controlling the various functions of the receiver.



Short News Items

The Institution of Electrical Engineers is to hold a conference at Savoy Place on 8 to 9 February on electronics design. The conference will be of a preliminary nature aimed at stimulating observation and experiment in this subject.

Arrangements for the conference are being made by an *ad hoc* committee set up by the I.E.E. Electronics Divisional Board to study fundamental processes and organizational procedures involved in the design of electronic equipment, to consider what can be done to help the electronic design engineer and to examine the problem of fostering design ability in educational establishments.

Further details and registration forms are available on application to the Secretary, I.E.E., Savoy Place, London, W.C.2.

The Institution of Electronic and Radio Engineers and the Electronics Division of the Institution of Electrical Engineers are joining together to hold a Symposium on "Microwave Applications of Semiconductors," at University College, London, from 30 June to 2 July, 1965.

The Symposium will cover fundamental properties and the development of semiconductor devices, their incorporation into complete microwave components and their systems applications. Applications to be described will include oscillators, harmonic generators, amplifiers, detectors, frequency changers, switches, and limiters, and systems made up of these component parts.

Offers of papers, with synopses, should be made as soon as possible to: the Secretary, Joint Organizing Committee for the Symposium on Microwave Applications of Semiconductors, The Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London, W.C.1, to whom requests for further information should also be sent.

The Ministry of Aviation, Horse Guard's Avenue, London, S.W.1, is to place on sale a bulletin entitled "Servomechanisms."

The bulletin will include abstracts from articles appearing in current journals and periodicals on control engineering and control systems, systems applications and instruments associated with servo-systems. It will include abstracts from reports originating in Britain and overseas.

The annual subscription of £3 will cover the 12 monthly issues plus an accumulated annual author and subject index. Single copies of the bulletin will cost 5s. and single copies of the annual index 6s.

The first number will be issued in January, 1965.

The Department of Electrical Engineering and Electronics, Brunel College, Acton, London, W.3, is to hold a course of 12 meetings beginning on 12 January, on statistical signal theory.

The course is intended for engineers and scientists who already have a basic knowledge of Information Theory. The topics covered will include: Deterministic signals, Fourier series and integrals, Sampling principles and the representation of signals by geometric spaces, Random signals, Stochastic processes and time series, Autocorrelation functions, Power spectra and Signal detection.

Further details are available from D. W. Lewen, Organizing Lecturer at Brunel College.

Transistorized submerged repeaters have recently been inserted in a submarine cable between St. Margaret's Bay and La Panne in Belgium, which increase its capacity from 216 to 420 telephone circuits. These are the first submerged repeaters using transistors to be used in an international submarine telephone cable and the first time that so many circuits have been operated on such a cable.

The cable was laid in 1948 and is of unusual design in that the polythene dielectric incorporates an air space obtained by a helix of polythene cord laid over the centre conductor before extruding the main polythene dielectric. The coaxial pair has an internal diameter of 1·7in for its outer conductor. The route length is about 47·5n.m.

The 216 telephone circuits obtained on the cable before the repeaters were inserted were transmitted from the U.K. to Belgium in the band 60-956kc/s and in the band 1152-2048kc/s in the reverse direction. At La Panne the 216 circuits were connected at line frequency range (60-956kc/s) to two coaxial tubes of a conventional coaxial land system to Brussels. The new 420 circuit system uses line frequencies in the bands 312-2044kc/s and 2544-4276kc/s, and, as for the earlier system the circuits are extended to Brussels in the line frequency range 312-2044kc/s over the coaxial land system. The highest working frequency over the submarine cable is about 4·3Mc/s at which frequency the total cable loss is about 130dB. The two submerged repeaters amplify the two directions of transmission using common amplifiers and have a gain of about 45dB at 4·3Mc/s. They are inserted in the cable at approximately 16n.m. and 32n.m. from the La Panne terminal.

The terminal transmission and power feeding equipment are fully transistorized and were manufactured on behalf of Submarine Cables Ltd by the Telecommunications Division of Associated Electrical Industries Ltd, one of its parent companies. Two pilot frequencies located near the edges of the traffic band in each direction of transmission continuously monitor the system performance. Variable equalizers are provided at the terminal stations to maintain the performance with changes of cable attenuation due to seasonal variations of sea temperature.

A three-way international radio, telephone and telecast cable 2 880 km long and linking East Germany, Russia and Czechoslovakia, began operating at the end of 1964.

The cable links Moscow with Kiev, Katowice, Brno, Prague and Berlin. Up to 1 920 telephone calls and two television transmissions can be handled simultaneously.

The new link, the longest in Europe, puts Moscow in direct contact with Berlin. It is the result of conferences held in Brno in May, 1962, and in Kiev in April, 1964. A recently developed Russian transmitting system is being used for the television and telephone links.

BINDING OF VOLUMES

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

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

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

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

The following are also available from our Circulation Dept.:

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

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

The total exports of valves, tubes and semiconductor devices during the third quarter of 1964 amounted to nearly £2½M according to figures based on the Customs and Excise returns issued by the British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA).

This is slightly down on the previous quarter, but there was an increase in some of the individual categories, such as microwave tubes, photo electric cells; stabilizing and cold cathode valves.

A Redifon G142 80 watt radio beacon transmitter operating at a frequency of 346·5kc/s has now been installed at Riverside Park, Dundee, providing an omnidirectional navigational aid for aircraft using the Riverside Airstrip.

The G142 covers a frequency range of 200 to 420kc/s and complies fully with ICAO requirements. It is used as a route marker, an approach or 'holding' aid in the vicinity of an airfield and it is also employed as a locator beacon in instrument landing systems. The output power of the beacon is controlled in steps down to one sixteenth of full power.

A three-company team representing interests in Britain, France and Germany, has been asked by the European Space Technology Centre to prepare a design study for a 35ft space simulation chamber. The Electronics Group of Associated Electrical Industries Limited has been commissioned to join W. C. Heraeus of Germany and Societe Generale du Vide, France, in a design study to show whether the installation of such a large chamber is essential for the ESTeC satellite test and launch programme. It is estimated that a 35ft chamber would cost around £2M.

The same companies have also received an order from ESTeC to design and build a smaller space simulation chamber of 15ft diameter. This will simulate the thermal conditions to which a satellite is exposed, and will be provided with equipment to simulated radiation, pressures and temperatures in space. It will also simulate the orbital motion of satellites.

The European Space Technology Centre in Delft, Holland, is a major establishment of the European Space Research Organization with responsibilities for Europe's satellite programme. It has a budget of £100M to be spent over the next eight years, subscribed to by all major European governments in the West. Britain, France and Germany have been the largest subscribers.

ESTeC's terms of reference include the design and testing of satellites for research purposes, launching, tracking and the collection of scientific data.

AEI's interest in space simulation now includes development work on a solar simulator to provide radiation similar to that from the sun. The power requirement for this "artificial sun" would be 50kW for the 15ft chamber, rising to over 1000kW for any 35ft simulator. The AEI Electronics Group is also engaged in the development of simulated earth conditions, and satellite orbital motions.

A revision of B.S. 2475 has recently been carried out and is now published by the British Standards Institution under the changed title of 'Octave and one-third octave band pass filters'.

Two main reasons lie behind this revision. An improved method of calculating the loudness of a noise has been evolved since the standard was first published in 1954, and this requires more detailed information on the composition of the noise than is given by an octave band analysis. One-third octave band filters have now been developed and these are included in the new edition. This more detailed information also permits noise reduction measures to be planned more accurately.

In addition, the publication of B.S. 3593 'Recommendation on preferred frequencies of acoustical measurements' has entailed bringing the mid-frequencies of the bands into line with the preferred values given in that standard.

The new edition of B.S. 2475 is in line with a specification now being prepared by the IEC (International Electrotechnical Commission) except that it omits half-octave filters.

Copies of B.S. 2475 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 5s. each. (Postage will be charged extra to non-subscribers).

An order worth £35 000 for 100 electronic measuring instruments of a particular type has been received from China by Burndept Electronics Ltd, Riversite Building, Erith, Kent.

The instrument has applications in hospital radiology, and as a safety device in the manufacture of luminous dials and switches and luminous paint. It can also be employed to check safety standards in measuring the flow of liquids and gases and in detecting faults in the production of special purpose steel, shipbuilding, bridge building, and tunnelling.

Thorn Parsons Company Ltd of Wellington Crescent, New Malden, Surrey, is now manufacturing a wide range of Positive Temperature Coefficient Resistors (p.t.c.r.). These resistors are made from barium titanate and have the characteristic that below the Curie temperature the resistance remains approximately constant while above this temperature the resistance increased sharply.

This makes the resistor suitable for excess temperature detection with the added advantage that the unit incorporating it will automatically fail safe should the resistor or its associated circuits become open circuit.

P.T.C.R. devices with varying coefficients can be produced with extremely sharp resistance changes over selected temperature ranges, or smaller resistance changes over larger temperature ranges. Standard p.t.c.r.'s have a tolerance on room temperature of ±10 per cent but devices with ±5 per cent tolerance can also be supplied.

The p.t.c.r. can also be used as a constant current device over a considerable voltage range. This current can be controlled at a selected value from a few milliamperes to approximately 100mA.

Ferranti Ltd has set up a new department called the Scientific Instrument Control Department, within the Ferranti Numerical Control Division. Its main interest is in the control of the position of special scientific instruments to a high degree of accuracy, using moire fringe techniques in the linear and circular modes. The scope of the work is broad, ranging at present from the positioning of crystals used in X-ray and neutron crystallography to the control of a large radio telescope for the Observatoire de Paris at Nancy. Diffraction work is being done in conjunction with Messrs. Hilger & Watts and the U.K.A.E.A. Harwell.

The Department is located at Dalkeith, Midlothian.

Rohde & Schwarz of Munich, Germany, has received an order from the German meteorological service for a v.h.f. receiving system for picture transmissions from weather satellites. This system which is to be set up in Offenbach will enable pictures of cloud structure taken by weather satellites to be observed over a large area and to be evaluated immediately for the investigation of weather conditions.

The new receiving system has a punched tape controlled antenna and will be the first in national use in the Federal Republic.

The experimental dual-transmitter stereophonic transmission of the BBC have now been closed down. The BBC has already done a great deal of experimental work and is collaborating with the broadcasting organizations of other European countries and with the European Broadcasting Union in investigating stereophonic systems. The question will be considered again at the Interim Meetings of CCIR Study Groups to be held in Vienna from 24 March to 7 April, 1965.

Meanwhile, the experimental stereophonic transmissions from Wrotham on 91·3Mc/s, using the compatible pilot-tone system, will be continued for the benefit of the radio industry. These transmissions have now been transferred from the mornings to the afternoons of Tuesday, Wednesday and Thursday of each week, the new times being as follows:—

2.30 p.m. to 3 p.m. Test tone transmission.

3.15 p.m. to 3.45 p.m. Programme test material.

Further consideration will be given to those tests for the radio industry in March next year when the Third Network Music Programme will be further extended into the afternoon period.

The Post Office will soon try out the pulse code modulation system over junction cables of 10-25 miles length which will provide 24 telephone circuits on two pairs of wires.

In this new system the speech information on each circuit will be sampled 8000 times per second, converted to a binary code and transmitted in the form of pulses at the rate of 1½ million per second. The pulses will be amplified every 2000 yards along the cable by regenerators housed in metal cases located in manholes. Power to operate the regenerators will be supplied from the terminal stations over the cables.

Trial systems are being provided by three contractors, Standard Telephones and Cables Ltd; Automatic Telephone and Electric Co. Ltd; and General Electric Co. Ltd. If tests now being carried out prove that such p.c.m. systems are acceptable, the Post Office will prepare a specification so that systems can be purchased and installed towards the end of 1967. The trial systems are being provided between Guildford and Haslemere (STC), Reading and Marlow (ATE), and Coventry and Rugby (GEC).

LETTERS TO THE EDITOR

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

Signal Flow Graphs

DEAR SIR,—Mr. Wheaton's letter in the issue of July, 1964, indicates that the signal flow graph technique breaks down in an attempt to perform the parameter inversion of transistor parameters. A careful study, however, reveals that this is not really the case.

In order to arrive at the required result, a different node inversion has to be performed as shown in the signal flow graphs, Fig. 1.

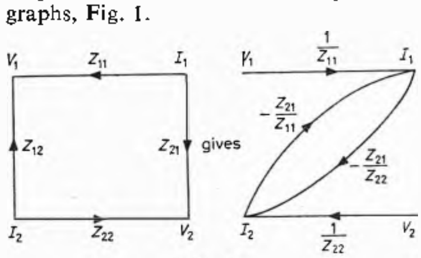


Fig. 1

The second signal flow graph can then be reduced to obtain either I_1 or I_2 by elimination of nodes I_2 *or* I_1 respectively, so that:

$$I_1 = \frac{Z_{22}}{Z_{11}Z_{22} - Z_{12}Z_{21}} V_1 - \frac{Z_{12}}{Z_{11}Z_{22} - Z_{12}Z_{21}} V_2$$

$$I_2 = -\frac{Z_{21}}{Z_{11}Z_{22} - Z_{12}Z_{21}} V_1 + \frac{Z_{11}}{Z_{11}Z_{22} - Z_{12}Z_{21}} V_2$$

Yours faithfully,

D. S. RANE,
Pennsylvania Military College,
U.S.A.

DEAR SIR,—Mr. O. J. Wheaton suggests (letter, page 486, July, 1964) that signal-flow-graph methods cannot be used to transform Z-parameters to Y-parameters. This can be done by standard techniques[1] as follows.

Suppose the Z-parameters are given (Fig. A). It is permissible to invert any path originating at a source node. In particular, this applies to Z_{11} (form I_1) and Z_{22} (from I_2). In Fig. B, Z_{11} is shown inverted, and, as I_2 is still a source node, Z_{22} can be inverted (Fig. C).

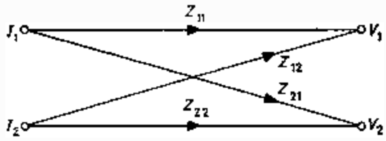


Fig. A

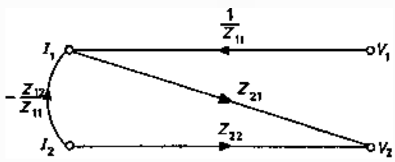


Fig. B

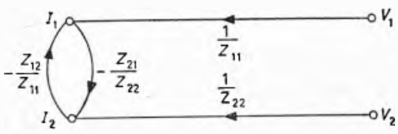


Fig. C

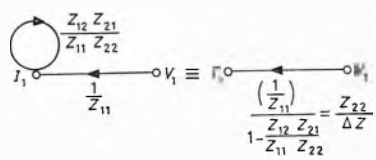


Fig. D

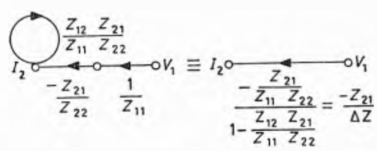


Fig. E

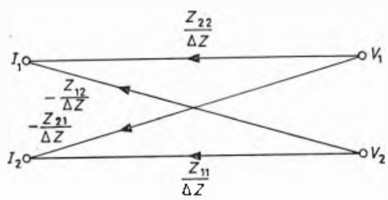


Fig. F

Considering now signal-flow from V_1 to I_1, this is as indicated in Fig. D. Signal-flow from V_1 to I_2 is indicated in Fig. E. In this way the complete graph is built up (Fig. F).

Yours faithfully,

A. C. DAVIES,
Northampton College of
Advanced Technology,
London.

REFERENCE

1. MASON, S. J., ZIMMERMANN, J. Electronic Circuits, Signals and Systems. Ch. 4, pp. 115-120 (Wiley, 1960).

The correspondent replies:

DEAR SIR,—I must apologise for wrongly concluding (in my letter, page 486, July 1964) that the transformation between Z and Y parameters cannot be achieved by the signal flow graph technique. Further, I must thank Mr. Davies and Mr. Rane for presenting the solution.

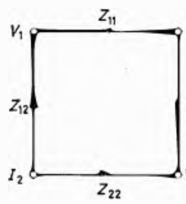


Fig. 1

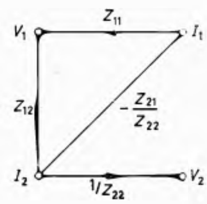


Fig. 2

Another interesting variation is the transformation between Z and h parameters. This can be achieved by the single inversion as shown in Figs. 1 and 2.

If Fig. 1 represents the Z parameters, then the inversion of the transmission Z_{22}, gives Fig. 2. In Fig. 2 the sources present are I_1 and V_2, which are those required for the hybrid parameters. Thus by inspection from Fig. 2,

$$V_1 = \left[Z_{11} - \frac{Z_{21} \cdot Z_{12}}{Z_{22}} \right] I_1 + (Z_{12}/Z_{22}) \cdot V_2$$

and

$$I_2 = -(Z_{21}/Z_{22}) \cdot I_1 + (1/Z_{22}) \cdot V_2$$

hence

$$h_{11} = Z_{11} - \frac{Z_{21}\,Z_{12}}{Z_{22}} = (\Delta Z/Z_{22})$$

$$h_{12} = Z_{12}/Z_{22}; \quad h_{21} = -(Z_{21}/Z_{22})$$

$$h_{22} = 1/Z$$

Yours faithfully,

OLIVER J. WHEATON,
Loughborough College of Technology.

BOOK REVIEWS

Introduction to Semiconductor Devices

By M. J. Morant. 126 pp. Demy 8vo. George G. Harrap. 1964. Price 18s.

THIS monograph covers all those aspects of semiconductor device physics considered essential to the electronics engineer if he is to fully realize the circuit potentialities of semiconductor diodes and transistors and, at the same time, be aware of their circuit limitations.

The discussion begins with a consideration of energy level diagrams in pure and extrinsic semiconductors. Then the lifetime and diffusion of excess carriers, the continuity equation and its implications are carefully explained. By now sufficient physical principles have been developed to undertake a fairly detailed study of the pn junction, and its behaviour in practice. Next the junction transistor is considered and its various d.c. characteristics are explained in physical terms. This is followed by a brief discussion of the various low frequency transistor equivalent circuits, high frequency effects, and the important methods of transistor fabrication. Finally, the operations of other semiconductor devices, e.g. tunnel diode, silicon controlled rectifier, field effect transistor, the Hall effect, etc., are briefly studied. Some 35 problems with answers are also provided.

The only criticism of the book is the lack of consistency in defining the directions of terminal current flow in the transistor. Thus, under d.c. conditions, the collector and base currents are shown flowing out of the device, whereas under a.c. conditions all terminal currents are shown flowing into the device.

The monograph is well written and is considered to be good value for money. It is unreservedly recommended to the undergraduate electronic engineering student and the practising electronic engineer seeking a first introduction to the physics of semiconductor devices.

S. S. HAKIM

Electronics and Instrumentation

By Robert L. Ramey. 321 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 55s.

FOR those intending to specialize in electronic instrumentation it is desirable that a good grounding in basic electronics should first be acquired, preferably outside a book on instruments. In the present book the author provides this grounding, but little on the instruments themselves. As stated in the preface, the text may be used in a first course for physics and electronic majors. This claim is reasonable for one chapter only is directly concerned with electronic instruments and this comprises about 10 per cent of the whole work.

The initial chapter deals in a summary manner with electron physics and the second with conduction in the solid state. Much of the latter chapter is concerned with basic electrical elements such as inductance, capacitance, etc., as well as semiconductors. The third chapter deals with electric circuit analysis based on the methods of Fourier, Kirchhoff, Thévenin, Norton and complex algebra. An account of topological analysis is given. Passive measurement systems form the subject of the fourth chapter, the systems referring to electromechanical instruments and bridges. Vacuum tube and transistor amplifiers are the subjects of chapters five and six, respectively, a large part of each of these two chapters being given to descriptions of the physics of vacuum tubes and transistors. Chapters seven and eight are respectively given to amplifiers, the latter to the feedback type.

In view of the very large number of electronic instruments in existence it is apparent that the ninth chapter of 33 pages on this subject can be no more than a summary. However, within the limits the space allows, a number of basic instruments are clearly described.

The tenth and final chapter is a useful one on the subject of transducers, but, again, the treatment is necessarily somewhat superficial.

Generally the book is well written and produced and is suitable for those requiring a work on those topics of basic electronics likely to be valuable in a subsequent study of electronic instruments. Each chapter concludes with a list of problems to which, however, no answers are provided.

F. G. SPREADBURY

Physical Electronics

By G. F. Alfrey. 220 pp. Med. 8vo. D. Van Nostrand. 1964. Price 27s. 6d. (paper bound), 50s. (cloth bound)

THE author states in his preface that this book is "... an attempt to provide ... introduction to the physical principles governing the operation of electronic devices. It is written for electrical engineers and for physicists ..." To that end the book deals very briefly with the electronic structure of atoms, the forces between atoms, emission of electrons, electron optics, control of electron flow in vacuum, conduction in gases, applications of gaseous plasma, electronic conduction in solids, semiconductor theory leading to pn junctions and transistors, magnetic properties including ferrites, dielectric materials, electrical noise and finally masers and lasers, all that in less than 200 pages—a brave attempt and perhaps not surprisingly not very successful within the author's terms of reference. Physicists are almost certain to find the physics contents too elementary. One would hope that it will be too elementary for most part even for electrical engineers—certainly the treatment of applications will be found at unacceptably low level for any electrical student and even the most elementary modern textbook on valves and transistors will devote more space to the physical principles than can be found in this book.

Among the curious inconsistencies one can mention the six pages devoted to vacuum valve characteristics without mentioning valve equivalent circuits, and not even a single drawing of transistor characteristics, but a somewhat indigestible paragraph on transistor equivalent circuits. One would also like to enquire how the author squares up the dimensions in his suggested expression for electromagnetic field $P = E + iH$ (page 26) and what happens when either E or H are complex in themselves. He blames tradition for not using this expression, but one wonders . . .

Nevertheless this book can be recommended to at least two classes of readers: those too old to have met modern electronics in their studies and those too young to have yet embarked on such studies. It is a useful popular exposition of the many aspects of the subject.

M. H. N. POTOK

The Use of Ferrites at Microwave Frequencies

By L. Thourel. 100 pp. Med. 8vo. Pergamon Press. 1964. Price 45s.

THIS book, Volume 26 in an international series on Electronics and Instrumentation, provides an introduction to the behaviour and applications of ferrites at microwave frequencies. The absence of detailed mathematical treatment brings the text within the grasp of the non-graduate electronic engineer who, finding himself employing ferrite devices from time to time, wishes to understand something of the principles of their operation.

Nearly one half of the text is devoted to a review of magnetism and the magnetic properties of ferrites couched, rather surprisingly, in the language of the physicist. The remainder of the text is concerned mainly with a description of resonance isolators and phase-shift circulators, although other components, including field-displacement devices, are mentioned.

There are two shortcomings with the text, the first is the early date of the original (1962), the present edition being a 1954 translation from the French. The text, in fact, reflects the state of the subject mainly before 1960, although a few specific references are dated in that year. As a consequence, the important junction-circulator device receives only slight mention and non-linear devices, such as limiters and ferrimagnetic delay-lines, do not feature at all in the book. The second shortcoming is the price as compared to the length. The body of the text comprises only 95 pages and for very nearly the same price several comprehensive texts are available which are three times the length. Thus this book will only appeal to the reader who finds the more comprehensive treatment beyond his grasp. To those persons the text can be recommended.

P. J. B. CLARRICOATS

Electrical Circuits Including Machines

By A. Draper. 289 pp. Med. 8vo. Longmans Green & Co. Ltd. 1964. Price 55s.

GENERALIZED machine theory provides a means whereby rotating machines may be considered as stationary networks. The solution of many machine problems is then reduced to linear circuit analysis. The first half of Mr. Draper's book provides the power engineer with more than an introduction into the circuit theory required for the solution of these problems. The chapters on matrix algebra, Laplace transforms and symmetrical components are particularly attractive, and the treatment of the two and three winding transformers in terms of their impedance matrices is to be commended. Other topics covered include an introduction to transformation theory and the use of the connexion matrix, mesh and nodal analysis, duality, two port networks, network theorems and mechanical analogues.

The second half of the book presents the basis of the generalized theory of machines, and in particular deals with the steady state analysis of the single and polyphase induction motors, capacitor motor, amplidyne, and synchronous machine. Equivalent circuits are derived from symmetric impedance matrices for the induction machines, and the torque expressions are obtained from the expression i_tGi. It is felt that a derivation of the steady state torque equation of a synchronous machine could have been included to some advantage. The section on the symmetrical short-circuit of an alternator may prove difficult for the student reader, particularly in the assumptions made and the hidden steps from one equation to the next.

It is very difficult to criticize such an excellent text, in which worked examples are given throughout and the material presented in a lucid and attractive manner. There are, however, text books on Generalized Machine Analysis which give a fuller treatment of the theory, but few include such a worthwhile exposition of the circuit theory required.

R. W. WHITEHEAD

Servomechanism Fundamentals and Experiments

By Members of the Staff, Philco Technological Center. 243 pp. Med. 8vo. Prentice-Hall International. 1964. Price 64s.

THIS is a disappointing book. The preface says that it has been prepared for use by university students and automation engineers, among others. Both the title and preface are misleading. The book is for the most part an introduction to selected electrical servo components at a level suited to senior technicians. If this had been done well the title and preface could be forgiven. In fact the individual sections are not well written, nor is the book as a whole well balanced.

The first three chapters, over one-third of the book, are entirely devoted to a laboured elementary treatment of synchros and closely related devices. Chapters 4 and 5 are presumably the fundamentals of the title. As an example of the style, on page 96 we learn ". . . instability . . . occurs . . . when the controlled device (because of the inertia of the controlling device) does not go to the ordered position instantaneously . . ." The authors find simplicity, brevity and accuracy incompatible; they frequently lapse into long, confusing and sometimes quite misleading attempts at non-mathematical exposition.

To say that chapter 6 is the best in the book is not high praise. It does give, however, a simple qualitative explanation of some selected transducers, including gyros and accelerometers.

Valve and transistor amplifiers, magnetic amplifiers, computer amplifiers, thyratrons, Ward Leonard sets and amplidynes are dealt with (in that order) in the 22 pages of chapter 7. The treatment is brief, rather confused, and contains errors. Chapters 8, 9 and 10 deal with servo motors, representative servos and maintenance techniques in a similar manner.

Some chapters are followed by a series of experiments for the diligent student. These use Philco apparatus, and it is obvious that the organization makes a great deal of interesting and useful tuitional equipment. It is a pity that this book is such a poor advertisement for it.

R. W. WILDE

Katalog normierter Tiefpassübertragungsfunktionen mit Tschebyscheffverhalten der Impulsantwort und Dämpfung (Catalogue of normalized low-pass filter transfer functions with pulse response and damping according to Chebyshev)

By Dr.-Ing. J. Joss. 140 pp. Royal 8vo. Westdeutscher Verlag. 1964. No. 1329. Carton covers DM. 43.-

Pulse shaping filters are used as transmitter filters in carrier frequency systems. As the calculation of these filters for individual application takes a long time, the normalized poles and zeros of appr. 900 of these low-pass transfer functions are given in catalogue form. The principles essential for an understanding are given together with instructions for the use of the tables.

Paramagnetic Resonance Vols. 1 & 2

Edited by W. Low. 921 pp. Med. 8vo. Academic Press. 1963. Price: Vol. 1: 114s. 6d., Vol. 2: 136s.

This two-volume work provides a comprehensive information source on major areas of interest in paramagnetic resonance. It contains about 180 papers dealing with the progress made in the field in the last 16 years.

The papers are primarily concerned with solids. The majority of contributions cover electron spin resonance in the iron group and the rare earth groups. Special attention is given to rutile, titanates, aluminium oxide, and calcium oxide. The work provides coverage of the theory related to these spectra and the effects of electric fields, pressure, and temperature on them.

Also provided are contributions on relaxation phenomena, electron spin resonance in semiconductors, and biological materials. Finally, there are a few selected papers on spin resonance of F centres, organic materials, and glasses.

Dielectrics

By J. C. Anderson. 171 pp. Crown 8vo. Chapman & Hall Ltd. 1964. Price 28s.

This book covers the syllabus for final-year undergraduate Electrical Engineering and Physics courses, together with more advanced treatments suitable for first-year postgraduate study. With its extensive bibliography the book will be of value to research workers in the field of dielectric materials.

Semiconductors

By H. Teichmann. 139 pp. Crown 8vo. Butterworth & Co. Ltd. 1964. Price 22s. 6d.

This textbook for general and advanced students bridges the gap between the purely technical literature and that which is theoretical and scientific. Little previous knowledge of the subject is assumed, and new information is linked to the fundamental aspects of physics and electrical engineering.

For the student who requires further information, a more detailed theoretical treatment is given in separate paragraphs which may be omitted on first reading; and there is a brief introduction to the essential points of wave mechanics, insofar as they are necessary to the understanding of semiconduction phenomena.

Analogue and Digital Computer Methods

By D. J. Harris. 112 pp. Demy 8vo. Temple Press Books. 1964. Price 30s.

In this monograph three types of computer, the analogue, the digital and the digital differential analyser, are examined in connexion with the solution of control system design problems in general and missile design problems in particular. The emphasis of the work is upon a comparison of computing techniques.

Other aspects of the subject which receive detailed coverage include: the general method of problem preparation for computer solution; the internal operation of each type of computer; and scaling methods and programming techniques as specifically applied to problems of missile design.

Computing Methods in Optimization Problems

Edited by A. V. Balakrishnan and L. W. Neustadt. 327 pp. Med. 8vo. Academic Press. 1964. Price 60s.

This book contains papers presented at the conference held at the University of California, Los Angeles in January 1964 and deals with topics such as hybrid computing methods, variational theory and optimal control theory, dynamic programming and invariant imbedding.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

(Voir page 59 pour la traduction en français; Deutsche Übersetzung Seite 66)

DIGITAL TAPE TRANSPORT

Plessey-UK Ltd, Control & Nucleonics Division, West Leigh, Havant, Hampshire

(Illustrated below)

The M3000 uses a new pneumatic-pressure tape drive principle which eliminates the disadvantages found with transports using pinch-roller drives. With vacuum reservoirs and a continuous reel servo the M3000 has a very high basic performance specification with high reliability at low cost, to meet the requirements of computer designers and users. It can also be supplied to meet a variety of lower performance specifications, as required.

Low skew is provided by the even distribution of acceleration forces across the tape; a characteristic of the pneumatic drive capstans. No electronic deskew is necessary.

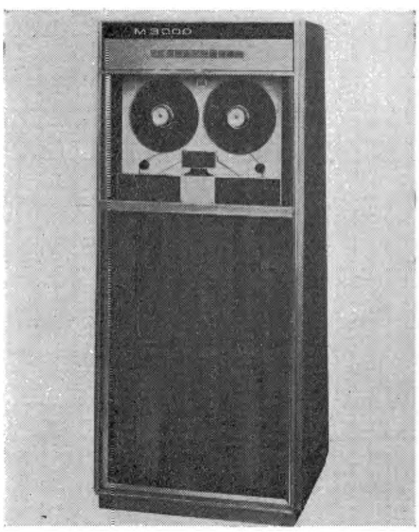
Gentle tape handling and reeling is a feature of the M3000 and overall tape wear is very low. This is due, particularly, to the low drive acceleration force, which is applied over a large area of the capstan and to the fact that there is no mechanical impact on the tape by the drive capstans. Tape acceleration is rapid and uniform, which keeps tape stresses to a minimum.

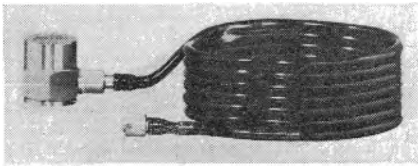
A very important requirement for high computer utilization is rapid interchangeability of tape reels. The M3000 provides this by straight line tape threading, together with a high speed rewind facility.

The M3000 was designed for high reliability, to provide low maintenance and downtime for the user. The pneumatic-pressure drive system, with its simplicity of design, has enabled these design objectives to be realized.

The M3000 uses only solid state electronics together with the latest techniques for stable operation over a wide range of supply voltage, ambient temperature and humidity. It is designed to utilize standard single-phase mains supplies; all other voltages required for the control and signal electronics are generated internally.

EE 77 751 *for further details*





ACCELEROMETERS

Distributed by: B. & K. Laboratories Ltd, Tilney Street, Park Lane, London, W.1

(Illustrated above)

These new Bruel & Kjaer high sensitivity accelerometers are designed for use in difficult environmental conditions and at high temperatures.

These accelerometers are small in size, light in weight, and very rugged in construction. They are suitable for the majority of vibration measurements, and particularly for applications involving severe environmental conditions such as humidity, high temperatures, magnetic fields, and corrosive atmosphere. Construction is sealed and waterproof.

Four different types are available, two with provision for water cooling, and top entry connexions, the other two having side entry connexions. A choice of frequency ranges is also available, the smaller types having titanium bases, and frequency ranges from 2 to 8000c/s (2 per cent) or 2 to 12000c/s (10 per cent), and the larger types with steel bases, having frequency ranges from 2 to 10000c/s (2 per cent) or 2 to 18000c/s (10 per cent).

Sensitivity of the smaller models is 10 to 20mV/g and the larger models 40 to 60mV/g, and each individual accelerometer is supplied with its own calibration chart.

All models have low transverse sensitivity, negligible mounting error, and can be operated at temperatures up to 260°C. The water cooled models can be operated at temperatures up to 1000°C.

Bruel & Kjaer accelerometers are supplied singly in a box, complete with accessories, or in packages of five, with connecting cables.

EE 77 752 *for further details*

INSTRUCTIONAL SERVO SYSTEM

Servomex Controls Ltd, Crowborough, Sussex

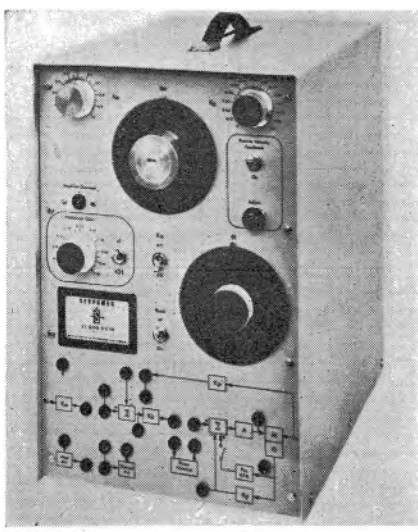
(Illustrated on right)

The instructional servo system type SS132 has been designed to fulfil the requirements for an instrument for teaching the fundamental principles of control engineering and which, while being reasonably cheap, would resemble closely the type of equipment which students would later encounter in industry. The teaching of high performance military servo systems is already catered for.

In order to fill this requirement Servomex Controls Ltd has designed the instructional servo system type SS132 which is a position control servo using 50c/s carrier throughout. It is provided with modulator and demodulator to allow it to be driven from a simple low-frequency oscillator and the results are shown on an oscilloscope or recorder. The controls are carefully calibrated so as to allow the student to carry out quantitative experiments which will adequately confirm the relevance of simple linear theory.

The instrument has distinct blocks with known transfer functions and adding units which only add. This is a considerable help to the student because he can concentrate on the analysis of the servo proper. An important feature is that all its units such as the summing and phase advance units are non-interacting. This enables the student to determine which component is performing a particular operation. In order to avoid unintentional saturation an overload indicator is provided which reveals amplifier overload by means of a flashing light. Positive velocity feedback can be switched in to eliminate viscous friction and to increase the general ease of measurement.

The motor drives the load through a coupling with marked compliance and a large inertia disk is provided which can be attached to either side of the compliant coupling or removed altogether. Also provided is a simple drum which



EE 77 122 *for further details*

VALVES FOR VHF COMMUNICATIONS

6146 Beam Tetrode
can be used as R.F. Power Amplifier up to 175 Mc/s or as an A.F. Power Amplifier or modulator.
Single Valve Class "C" Power Output at 175 Mc/s=25 Watts
I.C.A.S. Push Pull Class "AB1" Audio Power Output=120 Watts

6870 Special Quality VHF Power Pentode
for operation up to 150 Mc/s.
Anode Dissipation 6.3W max. Cathode Current 50mA max.

7558 Power Pentode
Operating up to 175 Mc/s for use as an R.F. Amplifier, Oscillator Frequency Multiplier or in A.F. Modulator and Amplifier applications.
Single Valve Class "C" Power Output at 175 Mc/s=6.5 Watts
Push Pull Class "AB2" Audio Power Output=20.5 Watts

Please ask for Data sheets

Thorn-AEI Radio Valves & Tubes Ltd.

HEAD OFFICE: 155 CHARING CROSS ROAD LONDON, W.C.2 Telephone: GERrard 9797

EXPORT DIVISION: THORN HOUSE UPPER ST. MARTIN'S LANE LONDON, W.C.2

Frequency range PM 5120
5 c/s – 600 kc/s (± 0.5 dB)
in 5 ranges
Error: < ± 2%
Distortion: < 0.5 % above 10 c/s

Frequency range PM 5121
1 c/s – 100 kc/s (± 0.5 dB)
in 5 ranges
Error: < ± 2 % ± 0.05 c/s
Distortion: < 0.5 % above 5 c/s

Output
Circuit: balanced or unbalanced
Amplitude: Max. 10 V into 600 Ω (balanced) or 20 V open circuit
Attenuation: 4 × 20 dB steps and continuously
Impedance: 600 Ω ± 2 %
Hum and noise: < 60 dB

Dimensions and weight
h × w × d: 29 × 22 × 32 cm (12 × 9 × 13 in)
Weight: 12 kg (26 lb)

The PM 5120 and PM 5121 are high-quality generators with low distortion and high stability over a frequency range extending from sub-audio frequencies to the medium wave radio-frequencies. Both generators have a constant output impedance and a highly stable output level, which can be obtained either balanced or unbalanced with respect to earth. As LF Generators are such basic tools for industry and research applications range from electronics to mechanics, from chemistry to ultrasonics, from medicine to geophysics.

PHILIPS

electronic measuring instruments

Sales and Service all over the world
For the U.K.:
The M.E.L. Equipment Company Ltd.,
207 Kings Cross Road, London WC1

can be attached to the output shaft and from which known weights can be suspended. By this means the student can measure the resistance of the servo to disturbing torques and relate it to the gain in the system.

A handbook written by a professional teacher of control engineering is supplied with each set. As well as giving details of the instrument it describes a number of experiments. One of these is the measurement of dead zone and its variation with change of gain.

EE 77 753 *for further details*

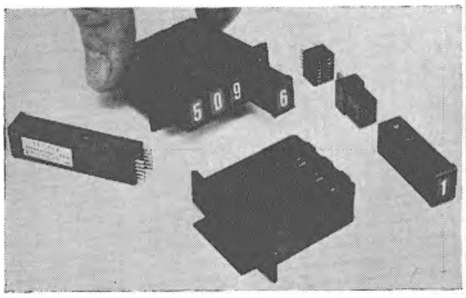
MICRO-INDICATORS

Distributed by: Kynmore Engineering Co. Ltd. 19 Buckingham Gate, London, W.C.2

(Illustrated below)

The Teldix model MA-11 micro-indicator, manufactured by Teldix, a German subsidiary of Telefunken and Bendix, it is now available through Kynmore Engineering Co. Ltd.

Each micro-indicator displays the figures 0 to 9 and one set of colour bars,



but special units can be provided to display letters, symbols, or any required sign. The white-on-black characters are 0·11in wide by 0·22in tall, and are projected on to a matt frontal screen mounted in a black plastic casing, from eleven individually beamed lamps in each unit. Overall dimensions of a single unit are 0·34in wide, 0·51in deep and 1·82in long. Cassettes are available to accommodate any number of units mounted side by side to make up decade or other displays. Standard black anodized light alloy cassettes are made for mounting up to six units, but as many as 400 units (50 × 80) have been mounted together for a digitized radar display.

The micro-indicator itself, i.e. the individual display unit less cassette, consists of three parts (a socket, lamp unit and display unit), all of which plug firmly together. In operation, the socket is normally screwed to the rear of the cassette, and the lamp unit and display unit are screwed together. In the event of lamp failure these two can be instantly unplugged from the socket, and withdrawn together from the front of the cassette, no tools being required. The display unit and replacement lamp unit are then reinserted in the front of the cassette, and again plugged into the socket by the miniature printed connector. The various units are suitably sprung to hold firmly and positively together.

Choice of logic circuits is left to the user, most types of transistorized decade logic being suitable. Operating from $1\frac{1}{2}$V at 70mA (±5mA for maximum lamp life), power consumption is only 0·1W. Characters are visible at angles of up to 30° from the normal in daylight. Nominal lamp life is 500 hours.

EE 77 754 *for further details*

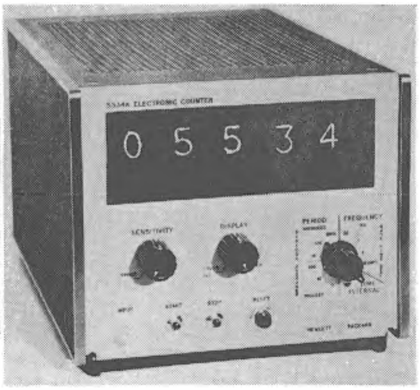
FREQUENCY COUNTER

Hewlett-Packard Ltd, Dallas Road, Bedford

(Illustrated below)

A new compact 2Mc/s, five-digit frequency counter, type 5534A, has been introduced by Hewlett-Packard Ltd.

Measuring facilities include single and multiple period average measurement in decade steps up to 10^5 periods, pulse operated time interval with either the 100kc/s time-base frequency of the instrument or any externally applied frequency up to 2Mc/s, frequency ratio or



multiples of ratio, and the counting of random events.

A clear, well-defined display of the counter reading is provided by cold-cathode numerical indicator tubes. Read-out storage provides a continuous display of the most recent measurement, even while the instrument is making a new measurement. Sampling time is, therefore, reduced to a minimum.

High sensitivity and a 1MΩ input resistance (shunted by 15pF) enables the instrument to detect and measure signals down to 100mV r.m.s.

Controls are simple and functional; the decimal point is automatically positioned and semiconductors are used throughout. The cabinet is suitable for bench or rack mounting and the plug-in printed board structure provides easy accessibility for instrument maintenance.

EE 77 755 *for further details*

DELAY LOOP RECORDER

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustrated above right)

This equipment has been developed by Gresham Lion Electronics Ltd for the recording and playback of audio signals or transients on a continuous loop of magnetic tape.

Three different lengths of loop may be used, 150in, 112in or 75in and, in the standard machine, two tape speeds are available, $7\frac{1}{2}$in/sec and $3\frac{3}{4}$in/sec.

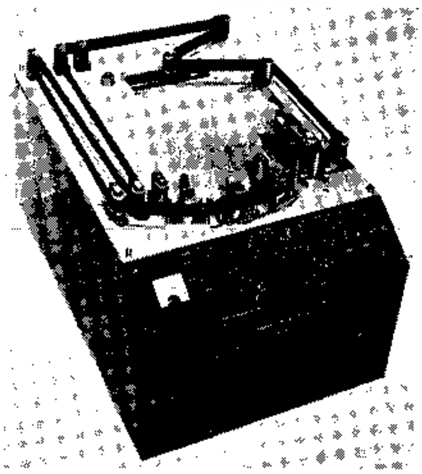
At the higher speed the three loop lengths give recycling times of 20sec, 15sec and 10sec, the corresponding figures for the lower speeds being 40sec, 30sec and 20sec.

Other tape speeds and delay times may be obtained by a simple change of capstan motor and the number of idler pulleys.

Fourteen channels on 1in tape are available, each with a transistorized plug-in record amplifier and playback amplifier, a switchable meter being provided to monitor record and playback levels.

A switch, cutting out the erase and recording circuits, enables a complete loop of recorded signal to be 'frozen' on the tape.

Due to the built-in azimuth accuracy



of the Gresham record playback anc erase magnetic heads, these are com pletely interchangeable without mechani cal re-adjustment.

EE 77 756 *for further detail*

A.C.-D.C. COMPARATOR

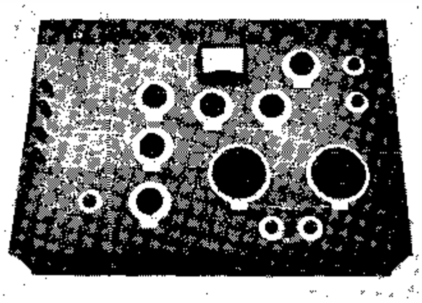
Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustrated on page 54)

This is a completely new, self-con tained, mains-operated instrument de signed to measure the r.m.s. values o alternating currents and voltages to ar accuracy of ±0·05 per cent over a fre quency range of 25c/s to 20kc/s.

A single vacuo-junction, having ar a.c. to d.c. transfer error of less thar 0·01 per cent, is used as a transfe element, and the circuit is compensate so that replacements can be fitted with out the need for recalibration.

An important advantage of the desig is that the unknown voltage or curren is rapidly balanced against an adjustabl d.c. potential, which can then b accurately measured at leisure using d.c. reference current calibrated to giv the true r.m.s. value of the applie signal. The reference current is derive



from a Zener stabilized supply unit having an output variation of less than 0·002 per cent for a mains voltage fluctuation of 10 per cent.

The only external units required are a standard cell and a suitable reflecting galvanometer, the instrument being otherwise self-contained with its own built-in potentiometer circuit. A useful feature, which reduces the probability of overloading the vacuo-junction, is a built-in meter that enables the approximate value of the unknown signal to be determined prior to setting the controls.

The instrument has eight voltage and current ranges, covering values from 0·5 to 300V and from 5mA to 3A respectively.

EE 77 757 *for further details*

A.C.-D.C. TEST SET

Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(Illustrated below)

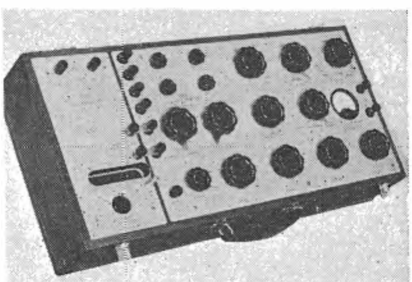
This is a compact versatile multi-purpose test set for the measurement of resistance, direct and alternating currents and voltages.

The instrument consists of a four-dial Rayleigh potentiometer, a.c.-d.c. thermal transfer standard, potential divider, current shunts, 50Ω spot reflecting galvanometer and all the necessary switches.

It is particularly useful for workshop or field use as it is portable and obviates the inconvenience of having to connect many components together which is normally associated with this class of measurement.

The instrument circuits can be supplied by two 2V accumulators or two 2V mains operated Zener diode supply units type P3/S.

It is simple to operate, having three switches for the selection of the voltage, current and resistance ranges; these switches are mechanically locked to ensure that the ranges not being used are returned to the 'off' position.



Provision is made for measuring currents higher than 5A by means of connecting shunt resistors external to the instrument.

A search meter is included in the input circuits to assist range selection when measuring voltage or current. For measurement of d.c. voltages up to 1·0999V, measurement is made direct with the potentiometer; above these voltages the potential divider is automatically included when range selecting. D.C. is measured with the potentiometer and current shunts; ranges being changed by altering the current range selector switch.

The a.c.-d.c. thermal transfer standard which is used for the a.c. measurements gives a true r.m.s. reading which is independent of waveform. The transfer device used is a vacuo-thermo-junction selected to have a low reversal error. The circuit is so arranged that the accuracy of measurement is entirely independent of characteristics of the vacuo-junction.

For resistance measurements one side of the Rayleigh potentiometer is used as the variable arm of a Wheatstone bridge. The bridge ratios are switch operated.

The instrument is contained in a teak case with a metal cover panel. The switches are "CROPICO" type S.P.1, which have a low contact resistance, positive indexing mechanism and high insulation. Comprehensive operating instructions are supplied with each instrument.

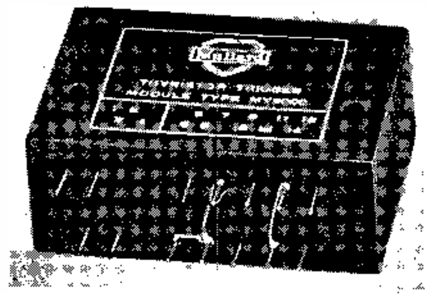
Forty ranges are provided covering resistance from 0 to 1MΩ, voltage from 0 to 600V d.c. and 2 to 600V a.c. (25c/s to 10kc/s), current from 0 to 5A d.c. and 5mA to 5A a.c.

EE 77 758 *for further details*

THYRISTOR DRIVE MODULE

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

(Illustrated above right)



Mullard Ltd has announced a thyristor drive module MY5000, which is designed to simplify the use of thyristors in a wide range of ordinary electrical applications. This is the simplest module of a range and will drive up to two 70A high current thyristors at the same time; for example, as two legs of a single-phase bridge circuit.

Although more complicated drive modules are necessary for sophisticated high accuracy motor speed control with complex feedback loops, this low-cost module will provide a satisfactory and economic solution for the many simple applications of motor speed control, furnace control, light dimming and also the control of a.c. power by a back-to-back thryistor arrangement.

One of the problems for thyristor users is the provision of reliable drive pulses that will satisfactorily fire the thyristors and also control the position of the pulse by means of a control element or feedback loop. Each application naturally requires its own feedback loop arrangement, but by using a Mullard module which provides properly toleranced thyristor drive pulses, only the d.c. and feedback problems remain. The more common of these problems will be covered in Mullard Technical Reports.

The trigger module MY5000 is completely protected against environmental effects by plastic encapsulation, a feature which protects circuit components and interconnexions from damage by mechanical vibration or shock.

EE 77 759 *for further details*

DELAY LINE STORES

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

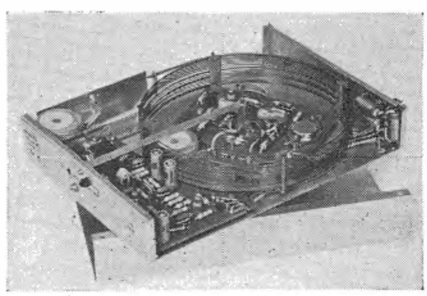
(Illustrated below)

Serial storage of information of up to 1 000 bits can be provided economically by a new range of ultrasonic delay lines in which all the piece parts are standardized.

Designated 'delay modules YL2108,' each unit in the range is supplied complete with magnetostrictive wire delay element and solid state input and output circuits. Any delay is available from 100μsec to 3·2sec, each type of unit being adjustable over a range of 20μsec.

The units will operate over a temperature range of −10°C to +60°C with bit rates of up to 350kc/s. A 12V power supply is required which may be positive, negative or symmetric. The units are therefore suitable for most types of both pnp and npn transistor logic systems, and are fully compatible with M.E.L. series 1 digital circuit blocks.

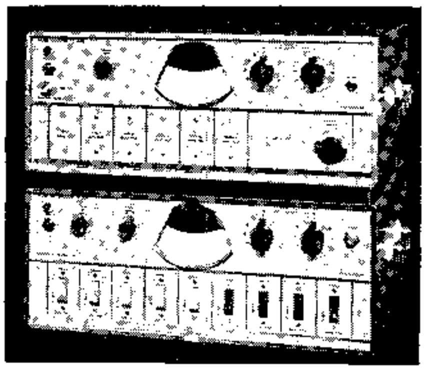
The delay lines normally operate in the cyclic mode in which the output is amplified and fed back to the input, the information being retained provided the power supply is kept on. The information may be extracted at any time, with a maximum access equal to the total delay in the line.



Many uses are envisaged in simpler kinds of data processing equipment. The lines can serve as temporary memories in small calculating machines, invoicing machines, on-line computers and similar applicatioss. They may also be used in conjunction with permanent memories such as core stores. The lines then retain information for short periods during calculations or sequencing operations.

The lines are contained in plated steel cases measuring 9¼ x 6¾ x 1½in. These may be mounted in any position or in the standard M.E.L. tray and rack system. One of the narrow sides of the unit forms the front panel on which an output test point and a delay control are provided. Power and signal input and output terminals are at the rear.

EE 77 760 *for further details*



WHITE NOISE TEST SET

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated above)

Marconi Instruments Ltd has introduced a new, fully-transistorized, white noise test set, type OA 2090, designed for testing cable and radio multi-channel links with capacities up to 2 700 channels, and meeting C.C.I.R. recommendations. It comprises two units and is considerably smaller in size and weight than its predecessors.

In multi-channel systems it is essential that inter-modulation interference in any channel, due to the traffic on other channels, is kept to a minimum. This crosstalk is produced mainly by non-linear and phase distortion and is audible as interference resembling random noise, which is evaluated by a noise power ratio measurement.

The test set operates on the following principles. White noise of the appropriate bandwidth and power level is applied to the system, simulating very closely a fully-loaded multi-channel telephone system. A filter with a narrow stop-band is interposed between the white noise source and the system, thus producing a quiet channel. A receiver tuned to the quiet channel is used to compare the noise level produced by intermodulation of the components of the white noise occupying the remainder of the frequency band with the original signal in the channel.

The noise generator uses a special semiconductor diode as the noise source. It will accommodate up to nine quick-change filter units—both band-limiting and the band-eliminating filters are incorporated. The filter units include switches with colour coded toggle-keys, and the positions in the circuit for all the filters occur before the power monitor; the noise-band limits and the slot positions are thus selected by simple switch operations, while the power level reference is continuously monitored. The system loading is adjustable by an attenuator at the output.

The receiver unit has provision for up to six spot frequencies and also employs inter-changeable filter units, which correspond to the generator band-stop filter frequencies. An input attenuator is so marked that the noise power ratio in decibels is read directly.

Dimensions of each unit are as follows: 7¼ x 18¼ x 17in, weight 26 lb. Rack mounting versions of the units are also available.

EE 77 761 *for further details*

FERRITE POT CORES

Standard Telephones & Cables Ltd, Magnetic Materials Division, Edinburgh Way, Harlow, Essex

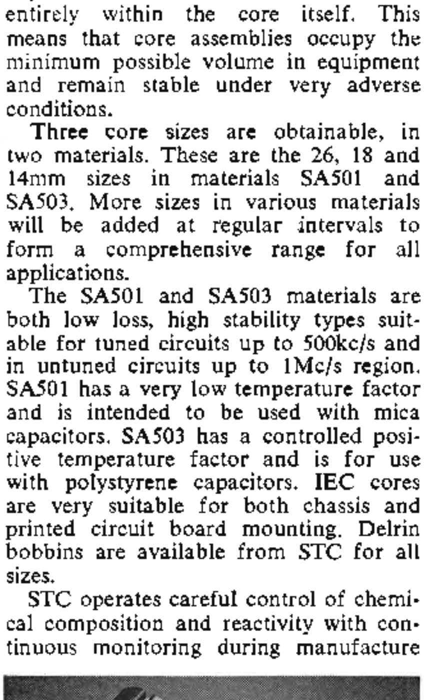
(Illustrated below)

STC Magnetic Materials Division has extended its range of high performance magnetic cores to include a new range of pot cores to international standards. A feature of this type of core is that the adjustment mechanism is contained entirely within the core itself. This means that core assemblies occupy the minimum possible volume in equipment and remain stable under very adverse conditions.

Three core sizes are obtainable, in two materials. These are the 26, 18 and 14mm sizes in materials SA501 and SA503. More sizes in various materials will be added at regular intervals to form a comprehensive range for all applications.

The SA501 and SA503 materials are both low loss, high stability types suitable for tuned circuits up to 500kc/s and in untuned circuits up to 1Mc/s region. SA501 has a very low temperature factor and is intended to be used with mica capacitors. SA503 has a controlled positive temperature factor and is for use with polystyrene capacitors. IEC cores are very suitable for both chassis and printed circuit board mounting. Delrin bobbins are available from STC for all sizes.

STC operates careful control of chemical composition and reactivity with continuous monitoring during manufacture



of IEC cores. Particular emphasis is on control of the ferrite microstructure and it is claimed that the process yields cores unmatched elsewhere.

EE 77 762 *for further details*

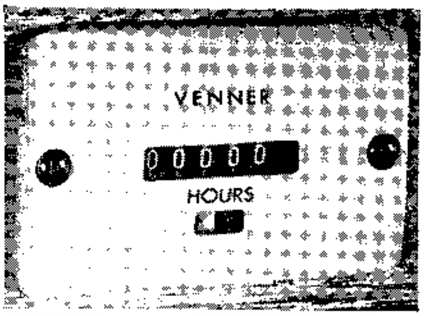
'HOURS RUN' INDICATOR

Venner Ltd, Kingston By-Pass, New Malden Surrey

(Illustrated below)

Venner Ltd has introduced a synchronous motor-driven hours-run indicator, type SHR1, for use with any apparatus operating on controlled frequency a.c. It reads in tenths of an hour up to 99 999·9 hours, with automatic restart at zero.

Applications include life tests, materials consumption, the running time of pumps and electronic equipment, and the indication of servicing and maintenance periods



on automatic lathes and similar factory machinery.

Two types are available, surface mounting with provision for cable or conduit entry, or flush mounting. A four-way terminal block allows for operation from three alternative voltages, 100 to 120V, 200 to 250V, 380 to 460V, 50c/s a.c. Consumption is 2·2W at 240W. Dimensions (flush mounted version) 3in wide, 2¾in deep, 2in high.

If required, the instrument can be supplied to register minutes instead of hours.

EE 77 763 *for further details*

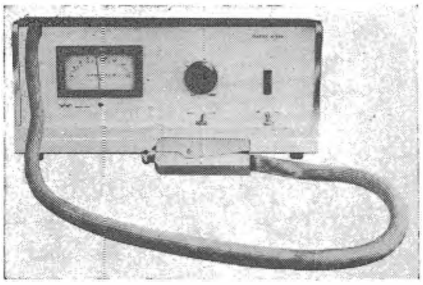
THERMO-PROBE

Distributed by: The Solartron Electronic Group Ltd, Victoria Road, Farnborough, Hampshire

(Illustrated on page 56)

The thermo-probe is a portable thermo-electric instrument designed to induce temperature variations on small components. Two thermo-electric cooling elements are mounted in a probe fitted with removable tip assemblies. Simple controls and a large temperature indicating meter are mounted on the front panel of a sturdy metal cabinet containing a d.c. power supply and heat exchange system.

The instrument will raise or lower the temperature of individual components while in circuit to assist prototype design and development. Probe tips of various shapes are available to provide maximum thermal contact with the component under test. Other possible applications



include goods inward inspection, trouble shooting and shrink fitting of small parts by direct heating or cooling. There will also be applications in chemical and medical research laboratories.

The temperature range of the thermo-probe is from −20°C to +70°C and the response time is two minutes to reach 60 per cent of final temperature.

Model TP-10 'thermo-probe' utilizes two thermo-electric cooling elements to pump heat between the test area and the ambient atmosphere. In the cooling mode, heat is extracted by specially designed tips from an object (e.g. an electronic component in an active circuit) and transferred by conduction to the faces of the thermo-electric elements. From here, the heat is pumped to the opposite faces of the elements by applied direct current and then conducted to the cooling liquid. The liquid is circulated through a liquid-to-air heat exchanger and the heat is transferred to the atmosphere. In the heating mode, heat flow is reversed by reversing the direction of the applied direct current. Heat, drawn from the atmosphere, is pumped into the test area. The quantity of heat pumped is a function of the direct current level. Temperature is sensed by a copper-constantan thermocouple connected to a Weston pyromillivoltmeter.

This instrument is manufactured by Daystrom Ltd of Canada and is available in the United Kingdom from The Solartron Electronic Group Ltd.

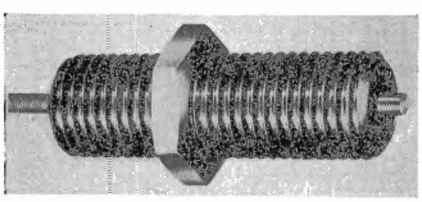
EE 77 764 *for further details*

MINIATURE MAGNETIC TRANSDUCER

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(Illustrated below)

Advance Controls Ltd has added a new miniature magnetic transducer type MT3SR to the existing range. These transducers contain no moving parts and will generate pulses from rotating shafts or gear wheels without any physical contact. Such pulses can be used to operate rotation monitors, speed switches, analogue or digital tachometers, ratiometers, speed difference indicators and for many other industrial control applications.



The transducer is mounted with its pole piece close to the path of moving gear teeth and with the recommended gap setting between 0·03 and 0·06in (allowing for some bearing play) pulses of more than 300mV amplitude are generated with peripheral speeds as low as 2in/sec.

This new transducer is hermetically sealed in a robust brass housing suitable for almost all industrial environments. It is 2in long and threaded $\frac{3}{8}$in B.S.F. for ease of mounting. Twelve other standard models for various applications are also immediately available and the special models which can be supplied number more than 60.

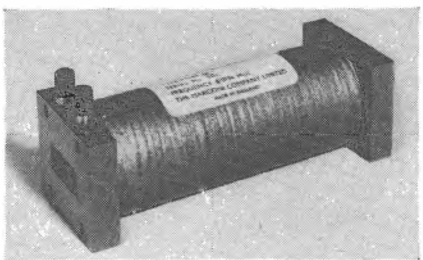
EE 77 765 *for further details*

WAVEGUIDE SWITCH

The Marconi Co. Ltd, Chelmsford, Essex

(Illustrated below)

A fully electronic waveguide switch with a power handling capacity of 2W,



covering the frequency range 7·75 to 8·5Gc/s, is now available from The Marconi Co. Ltd. The attenuation of the switch is 50dB, and it has been designed for use with microwave communications equipment, where it will provide considerable improvements in reliability, speed, space saving and efficiency over previous mechanical switching methods.

The new switch is in waveguide size No. 15 and consists of a length of specially fabricated waveguide containing a ferrite insert, with a coil wound round the outer body of the waveguide to produce a longitudinal magnetic field. When this field is energized, the attenuation rises quickly to 50dB. The switching time can be achieved in less than 1msec by a small transistorized switching circuit.

EE 77 765 *for further details*

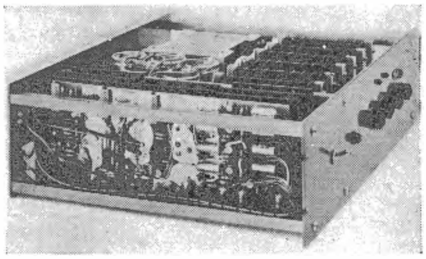
TUNNEL DIODE POWER SUPPLY

Coutant Electronics Ltd, 3 Trafford Road, Richfield Estate, Reading, Berkshire

(Illustrated above right)

The comparatively low voltages associated with tunnel diode circuits demand special techniques, particularly when the currents run into several amperes.

Coutant Electronics Ltd has designed a stabilized power unit to match into standard printed strip lines. Output connexions are made by soldering the



strip lines directly to the double-sided printed circuit in the power unit, with provision for ground plane continuity.

The output voltage is continuously variable between 0 and 100mV at a current of 8A and two auxiliary supplies of +5V and −5V at 6A are provided for associated transistor circuits.

Remote error signals are transferred from the load to the control amplifier through miniature coaxial cables and connexions. The response time of the stabilizer to step changes of load current approaches the theoretical minimum attainable, depending mainly upon the length of strip line employed between the power unit and the sensing point.

The power supply utilizes an all silicon semiconductor line up and will operate in ambient temperatures as high as 60°C. It is designed for standard 19in rack mounting and occupies a panel height of $5\frac{1}{4}$in.

EE 77 767 *for further details*

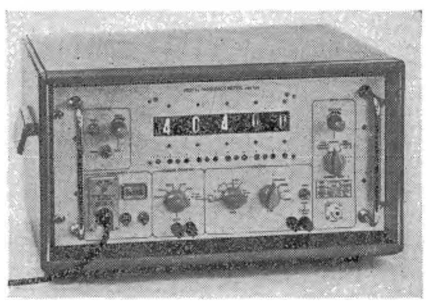
DIGITAL FREQUENCY METER

Dawe Instruments Ltd, Western Avenue Acton, London, W.3

(Illustrated below)

To meet the demand for a moderately priced digital frequency meter suitable for a wide range of general laboratory and industrial applications, Dawe Instruments Ltd has introduced the type 719A digital frequency meter. With a five-digit illuminated read-out and 10kc/s quartz crystal oscillator, it is suitable for a wide range of applications including frequency, period and speed (in rev/min) measurement, as well as random pulse counting.

The type 719A is based on normal digital counter-timer principles, the electrical input waveform being amplified and limited to provide a train of pulses which are counted during a known gating period, or are themselves used to control the gate. For frequency measurement, clock pulses from the crystal oscillator are used in conjunction with decade dividers to open and close the gate for



a preset period from 0·01 to 10sec, during which input pulses are counted and displayed. Speed is measured in the same way, over a gating period of 6sec, to provide the direct rev/min indication. For period measurement the procedure is reversed, the number of pulses from the crystal oscillator (or a decade division of this number) being counted during one cycle of the input. The instrument may also be used for 'stop watch' work, by counting the number of pulses over any fixed period of input. Random pulses may also be counted, provided the separation between successive pulses is not less than 0·1μsec.

Frequency may be measured from d.c. to 10Mc/s, period and random pulses up to 99999sec. Overall accuracy is ± 1 count, ± crystal accuracy, the crystal being set to one part in 10^6 at room temperature. The numerical display gives automatic decimal point and 'count exceeded' indication, the final display being visible for a preset period from 1sec to 15sec, or until reset manually. Transistorized construction is employed throughout, and a novel decade divider circuit (patent applied for) has considerably simplified the circuit.

The type 719A is suitable for a wide range of work, including tachometry, frequency calibration, relay contact timing, synchronized 'stop watch' work and phase and torque measurement. In particular, it will be suitable for many such situations where a digital frequency meter was previously considered too expensive.

Power consumption is only 15W (mains) and weight is 35 lb. Dimensions are 20½ x 11 x 12in deep, the case being suitable for rack or bench mounting.

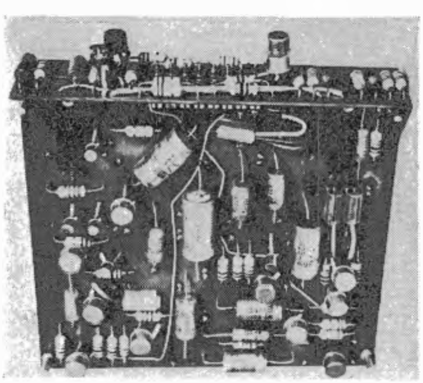
EE 77 768 *for further details*

OPERATIONAL AMPLIFIER

Bendix Electronics Ltd, High Church Street, New Basford, Nottingham

(Illustrated below)

Bendix Electronics Ltd has announced a new general purpose, solid state, chopper stabilized amplifier called the 470. The application of feedback round the amplifier produces a precise gain characteristic that is entirely a function of feedback impedances. In addition, the input is virtually at zero potential and no current flows into the amplifier, thus providing an ideal summing function for input currents. The 470 uses silicon epitaxial planar transistors for maximum reliability, and has a temperature range 0 to 70°C and a gain 8×10^6 at d.c. reducing to unity at 2Mc/s.



The auxiliary amplifier gain is 2000 reducing at 20dB/decade from 0·045c/s and the wideband amplifier gain is 4000 reducing at 20dB/decade from 1kc/s. The input impedance is 6kΩ and the output 10V into a load of 1kΩ.

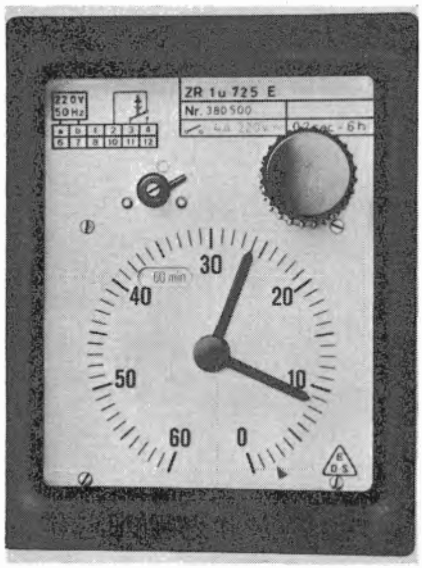
EE 77 769 *for further details*

PROCESS TIMERS

Distributed by Rayleigh Instruments Ltd, 271 Kiln Road, Thundersley, Essex

(Illustrated below)

A new range of high precision timers and time delay relays manufactured by DOLD & Sohne of Western Germany has recently been introduced into the United Kingdom by Rayleigh Instruments Ltd.



These instruments which are of the syncronous motor type and of well-proven design, are made to conform to very rigid timing and electrical specifications. They are available in various types, certain of which can accommodate up to five different time ranges. On multi-range instruments, range selection is by means of a simple control screw. Delay periods can be set anywhere between the limits of 0·1sec to 120h depending upon the range of the particular instrument chosen.

According to customers' requirements, these instruments can be fitted with contacts which will operate either at the beginning or end of the time cycle.

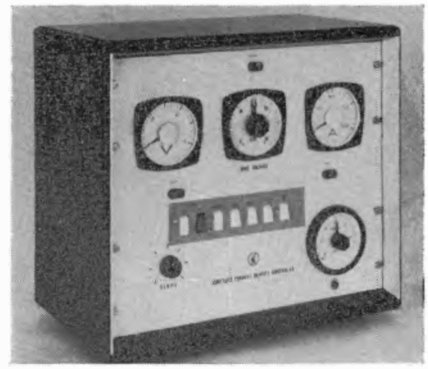
EE 77 770 *for further details*

AUTOMATIC POWER CONTROLLER

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated above right)

Kent Precision Electronics Ltd has produced a range of solid state automatic controllers of advanced design for use



with motorized regulators. Normal automatic control is achieved by switching motorized regulators to raise or lower the output as required.

The KPE range of instruments operates with a standard solid state calibrated control instrument for proportional control of a d.c. commutator motor. For large errors the motor is switched fully on and, within the proportional band of approximately 2 per cent, slows down to attain the set point with no overshoot or hunting. A dead space of less than 0·5 per cent is easily achieved in all normal applications. When conventional on-off control is required for the general commercial range of motorized regulators, a standard KPE instrument is available that uses a solid state calibrated control instrument which operates a standard relay for the on-off control function.

Programmed operation, as for example when constant current density control is required for electroplating vats, is available by the introduction of secondary operational inputs to the standard solid state calibrated control instrument. A standard constant current density controller is shown on the photograph.

The instruments are available as robust panel or rack-mounting units, and since there are no moving parts, automatic control systems using the KPE range of instruments feature reliable and maintenance free operation.

EE 77 771 *for further details*

AUTOMATIC ATTENUATOR

Sangamo Weston Ltd, Great Cambridge Road, Enfield, Middlesex

Sangamo International now has in production the automatic attenuator type SA-26. This device answers the need felt by those handling analogue data for almost every purpose where occasional peaks or unpredictable amplitudes overdrive the indicating or recording system —be it a magnetic tape recorder, oscillograph, chart recorder or plotter.

The model SA-2L attenuates signals above an adjustable trigger level by an accurately known amount and thereby widens the dynamic range of any constant noise system. The attenuator need not be limited to applications involving peaks. It can also be used as a range expander providing full system accuracy at two selectable points.

EE 77 772 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Electronics Division

Date: 4 January
Lecture: The Present State of Gallium Arsenide Technology
By: J. R. Knight

Date: 11 January
Lecture: Speech Compression
By: J. Swaffield

Date: 13 January
Lecture: Potential Levels in the Cental Nervous System—Their Detection and Significance
By: Sir Bryan Matthews
(Joint meeting with the Institution of Electronic and Radio Engineers)

Date: 22 January Time: 2.30 p.m.
Discussion: The Direct Recordings of Biological Signals
(Joint meeting with the I.E.R.E. Medical and Biological Group)

Date: 25 January
Times: 10.30 a.m., 2.30 p.m. and 5.30 p.m.
Second Colloquium on: Logic Circuits
(All wishing to attend must register)

Science and General Division

Date: 5 January
Lecture: Dynamic Performance of Induction Motors in Control Systems
By: J. C. West, B. V. Jayawant and G. Williams

Date: 19 January
Discussion: The Performance of Hill-Climbing Systems—The Gradient Estimation Problem
Opened by: K. C. Ng and J. D. Roberts

Date: 22 January
Colloquium: Network Analysis

Date: 26 January
Lecture: Electrostatic Generators
By: N. J. Felici

Date: 29 January
Lecture: Measurement of Earth Electrode Resistance with Particular Reference to Earth Electrode Systems Covering a Large Area
By: G. F. Tagg

THE INSTITUTION OF ELECTRICAL AND RADIO ENGINEERS

All meetings at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, unless otherwise stated.

Joint I.E.E./I.E.R.E. Meeting

Held at: The Institution of Electrical Engineers, Savoy Place, W.C.2.
Date: 13 January. Time: 5.30 p.m.
Lecture: Potential Levels in the Central Nervous System
By: Sir Bryan Matthews

Electro-acoustic Group

Date: 20 January Time 6 p.m.
Lecture: Acoustic Communication Underwater
By: B. K. Gazey and J. C. Morris

Joint I.E.E./I.E.R.E. Computer Groups

Date: 25 January Time: 2.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, W.C.2
Colloquium: Logic Circuits
(Advancec registration will be necessary for attendance at this meeting)

Radar Group

Date: 27 January Time: 6 p.m.
Symposium of short papers on: Enhancement and Absorption of Radar Radiation
Contributions include: Enhancement of the Radar Echoing Area of Gliders
By: R. S. Badcoe and S. R. Arbuthnot
Omni-Azimuthal Radio Echo Enhancer
By: J. Croney

Southern Section

Date: 19 January Time: 6.30 p.m.
Held at: Brighton College of Technology
Lecture: Design and Applications of a Fast Industrial Static Switching System
By: G. W. Pointin

Date: 14 January Time: 7 p.m.
Held at: Farnborough Technical College
Lecture: Attitude Control of the Skylark Sounding Rocket
By: J. F. M. Walker

South Western Section

Date: 20 January Time: 7 p.m.
Held at: University of Bristol Engineering Laboratories
Lecture: Some Aspects of Radioactive Particle Detection Techniques and Associated Circuit Design
By: J. R. Brown

South Wales Section

Date 13 January Time: 6.30 p.m.
Held at: Welsh College of Advanced Technology, Cardiff
Lecture: Parametric Amplifiers
By: T. Buckley

South Midland Section

Date: 29 January Time: 7 p.m.
Held at: North Gloucestershire Technical College
Lecture: Digital Storage
By: W. Renwick

Merseyside Section

Date: 20 January Time: 7.30 p.m.
Held at: Walker Art Gallery, Liverpool
Lecture: Radio Astronomy
By: R. C. Jennison

East Midland Section

Date: 28 January Time: 6.30 p.m.
Held at: Loughborough College of Advanced Technology, Loughborough
Lecture: Nuclear Instrumentation
By: R. B. Quarmby

North Eastern Section

Date: 13 January Time: 3 p.m.
Held at: Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle upon Tyne
Lecture: Field Effect Transistors and their Applications
By: D. S. den Brinker

West Midland Section

Date: 20 January Time: 7.15 p.m.
Held at: Wolverhampton College of Technology
Lecture: Microelectronics
By: J. W. Granville

Scottish Section

Date: 13 January Time: 7 p.m.
Held at: The University, Drummond Street, Edinburgh
Lecture: Quality and Reliability
By: F. Baxter

Date: 14 January Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Lecture: Quality and Reliability
By: F. Baxter

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 8 January
Lecture: Television Coverage of the Tokyo Games
By: J. E. F. Voss

Date: 21 January
Lecture: Video Tape for Studio Inserts
By: R. Firkin

PUBLICATIONS RECEIVED

CVS CONTINUOUS VOLTAGE STABILIZERS is the title of a leaflet published by The British Electric Resistance Co. Ltd, Queensway, Enfield, Middlesex, describing their new range of stabilizers.

ELECTRONIC CATALOGUE. Electrothermal Engineering Ltd have now published the 1964/65 edition of their 50-page illustrated catalogue which gives full technical details on a wide range of products. It includes information on the Series G101 Precistors, Environmental Test Chambers, Thermal Chambers, Temperature Controllers and other electronic instruments and components. Copies of this catalogue can be obtained by writing to Electrothermal Engineering Ltd, 270 Neville Road, London, E.7, and requesting Publication No. 414.

SOLARTRON DIGITAL DATA SYSTEMS gives an outline of the Solartron philosophy of building custom-built logging systems from standard modules. It describes how data logging can be applied to automated recording of results from a jet engine test bed, and continues with detailed descriptions, with illustrations of a few of the Solartron Logging Systems installed throughout the world. Requests for copies of this booklet should be addressed to the Publicity Manager, The Solartron Electronic Group Ltd, Victoria Road, Farnborough, Hampshire.

TYPE 8 CLOSED-CIRCUIT TELEVISION is the subject of an EMI Electronics Ltd publication now available. One hundred and thirty users of the Type 8 closed-circuit television camera are listed and details are also given of the many optional accessories which enable an elaborate network of cameras and receivers to be built up by stages from a basic one-camera installation. A detailed technical specification completes the booklet, which can be obtained from EMI Electronics Ltd, Instrument Division, Hayes, Middlesex.

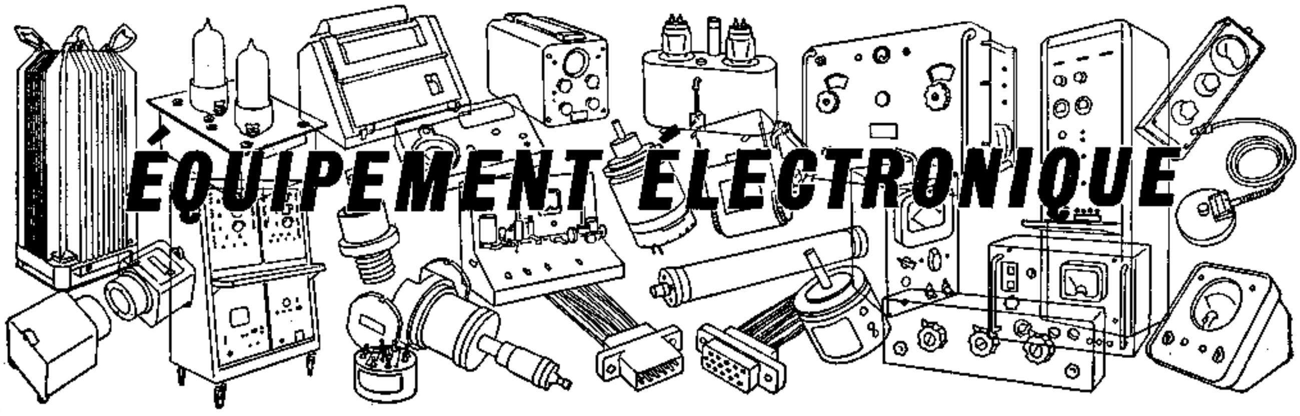
VIBRO-METER CORP. of Switzerland have recently opened a British office at Haletop Civic Centre, Wythenshawe, Manchester 22, for the distribution of their transducers and electronic instruments. The new office has been created to provide technical liaison between the factory in Fribourg and British engineers. Copies of a short form catalogue announcing the opening of the Vibro-Meter British office are available from their Manchester address.

FASTRAN DESIGN MANUAL gives details of the service offered to meet the requirements of the electronic industry for fast, reliable delivery of small power transformers. The service covers the supply of 50c/s, single-phase transformers up to 380VA output. Quantities of up to 25 will be despatched within 14 days of receipt of order. Further information on this service is available from the Assistant Sales Manager (Correx Department), Phoenix Telephones Ltd, Grove Park, London, N.W.9.

THERMOCOUPLES: THEIR INSTRUMENTATION, SELECTION AND USE by B. F. Billing, Senior Experimental Officer, Royal Aircraft Establishment, provides all the basic information required for the practical selection and use of thermocouples and the inspection of thermocouple installations. The properties of all thermocouples generally available are described, together with the principles and practical application of current sensitive and potentiometric instrumentation in manual and automatic forms. The price of this Monograph is 15s., postage 9d. and copies may be obtained from The Institution of Engineering Inspection, 616, Grand Buildings, Trafalgar Square, London, W.C.2.

STC SEMICONDUCTOR DEVICES DATA SUMMARY is the title of a new publication (No. M/106) on the ranges of semiconductor devices manufactured by ESC Components Group. This 46-page booklet gives the essential ratings, characteristics and outlines for: silicon and germanium diodes, silicon rectifiers, thyristors, silicon and germanium transistors as well as integrated circuits and other multiple devices. Copies are available from STC Components Marketing Division, Footscray, Sidcup, Kent.

NON-CUTTING USES OF INDUSTRIAL DIAMONDS is the title of a new publication now available, free of charge, from the Industrial Diamond Information Bureau, Arundel House, 36-43 Kirby Street, London, E.C.1.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 52 à 57*

ENTRAÎNEMENT DE BANDE NUMÉRIQUE

Plessey-UK Ltd, Control & Nucleonics Division, West Leigh, Havant, Hampshire

(Illustration à la page 52)

L'appareil M3000 utilise un nouveau principe d'entraînement de bande à pression pneumatique qui obvie aux désavantages des mécanismes à entraînement par rouleau à pince. Comprenant des réservoirs à vide et un rouleau continu, le M3000 a une spécification de base de rendement très élevée. Il est d'une grande fiabilité et d'un prix fort bas et répond donc aux besoins des réalisateurs et utilisateurs de calculatrices. Il peut être fourni pour répondre à une grande variété de spécifications de performance inférieures, selon les besoins. Une flexion en biais réduite est assurée par la distribution égale des forces d'accélération à travers la bande, ce qui est une caractéristique des cabestans à entraînement pneumatique. Aucun "débiaisage" électronique n'est nécessaire.

Le M3000 se distingue en outre par la douceur d'avance et d'enroulement de la bande, l'usure de cette dernière étant donc réduite. Cela est dû en particulier à la force d'accélération d'entraînement réduite et appliquée sur une surface étendue du cabestan, ainsi qu'au fait qu'aucun impact mécanique ne se fait sentir sur la bande par les cabestans d'entraînement. L'accélération de la bande est rapide et uniforme, ce qui réduit au minimum les contraintes imposées à la bande.

Une condition très importante de l'utilisation des calculatrices est l'interchangeabilité rapide des rouleaux de bandes. Cette condition est pleinement remplie par le M3000 grâce au filetage rectilinéaire de la bande et à la possibilité de rebobinage à grande vitesse.

Le M3000 a été conçu en vue d'une grande sécurité d'emploi. Son entretien se réduit au minimum. Le système d'entraînement par pression pneumatique, d'une grande simplicité de conception, a permis d'atteindre ces objectifs.

L'appareil comprend des composants constitués de corps solides et il s'inspire des techniques les plus récentes qui lui assurent un fonctionnement stable dans une gamme étendue de tensions d'alimentation, de températures ambiantes et de variations du degré d'humidité. Il est prévu pour l'alimentation secteur monophasée normale; toutes les autres tensions nécessaires à l'éléctronique de contrôle et de signalisation sont produites intérieurement.

EE 77 751 *pour plus amples renseignements*

ACCÉLÉROMÈTRES

Distributeurs: B. & K. Laboratories Ltd, Tilney Street, Park Lane, London, W.1

(Illustration à la page 52)

Ces nouveaux accéléromètres Bruel & Kjaer à haute sensibilité ont été étudiés pour l'emploi dans des conditions ambiantes défavorables et à des températures élevées.

Ils sont petits, légers et de construction très solide. Ils peuvent être utilisés pour la plupart des mesures de vibrations et, en particulier, dans des conditions ardues d'environnement telles que celles d'humidité et de température élevées, de champs magnétiques et d'atmosphère corrosive. Ils sont enfermés dans des boîtiers scellés et étanches.

Quatre modèles différents sont prévus: deux avec possibilité de refroidissement par l'eau et connexions d'entrée par le haut et deux autres avec connexions d'entrée latérales. Un choix de gammes de fréquences est également offert: les petit modèles étant à base de titane et à bandes de fréquences de 2 à 8000 Hz (2 %) ou de 2 à 18000 Hz (10 %) et les grands modèles étant à base d'acier et à bandes de fréquences de 2 à 10 000 Hz (2 %) ou de 2 à 18 000 Hz (10 %).

La sensibilité des petits modèles est de 10 à 20 mV/g et celles des grands nodèles de 40 à 60 mV/g. Chaque accéléromètre est fourni avec son propre diagramme d'étalonnage.

Tous les modèles ont une faible sensibilité transversale, une erreur de montage négligeable et peuvent être utilisés à des températures allant jusqu'à 260° C. Les modèles à renfroidissement par l'eau peuvent être employés à des températures allant jusqu'à 1000° C.

Les accéléromètres Bruel et Kjaer sont fournis individuellement dans un coffret complet avec tous les accessoires ou en lots de cinq appareils avec câbles de connexion.

EE 77 752 *pour plus amples renseignements*

SYSTÈME D'ASSERVISSEMENT D'INSTRUCTION

Servomex Controls Ltd, Crowborough, Sussex

(Illustration à la page 52)

Le système d'asservissement d'instruction type SS132 a été conçu pour répondre à la demande d'un instrument pour enseigner les principes fondamentaux de la technique du contrôle. Tout en étant raisonnablement bon marché, il ressemblerait étroitement au type d'appareil que les étudiants retrouveraient plus tard dans l'industrie. L'enseignement de systèmes d'asservissement militaires à grand rendement est déjà prévu.

Afin de répondre à cette demande, la société Servomex Controls Ltd a conçu le système d'asservissement d'instruction type SS132 qui est un système de contrôle de position utilisant une porteuse de 50 Hz. Il est muni d'un modulateur et d'un démodulateur ce qui permet de le mettre en action à partir d'un simple oscillateur basse fréquence, les résultats étant indiqués sur un oscilloscope ou un enregistreur. Les commandes sont étalonnées avec soin de manière à ce que l'étudiant puisse exécuter des expériences quantitatives qui confirment amplement la pertinence de théories linéaires simples.

L'instrument comprend des blocs distincts à fonctions de tranfert connues et des additioneurs qui se bornent à leur rôle. Cela constitue une aide précieuse pour l'étudiant car il peut ainsi se concentrer sur l'analyse de l'asservissement proprement dit. Une caractéristique importante est le fait que tous ses éléments, tels que les éléments de sommation et d'avance de phase, sont sans effet réciproque. L'étudiant peut donc déterminer le composant particulier qui effectue une opération donnée. Afin d'éviter la saturation involontaire, un indicateur de surcharge est prévu. Ce dernier indique toute surcharge de l'amplificateur au moyen d'un voyant lumineux. La réaction positive de vitesse peut être branchée afin d'éliminer la friction visqueuse et augmenter la facilité générale de la mesure.

Le moteur fait passer la charge à travers un couplage. Un grand disque d'inertie peut être fixé à l'un ou l'autre des côtés du couplage ou supprimé entièrement. Il y a aussi un simple tambour qui peut être fixé à l'arbre de sortie et auquel des poids connus peuvent être suspendus. Par ce moyen, l'étudiant peut mesurer la résistance de l'asservissement à des couples perturbateurs et la rapporter au gain dans le système.

Un manuel d'instruction, rédigé par un professeur de la technique du contrôle, est fourni avec chaque appareil. Tout en donnant des détails de l'instrument, il décrit aussi certaines expériences. L'une d'elles concerne la mesure de la zone morte et ses variations en fonction des changements de gain.

EE 77 753 *pour plus amples renseignements*

MICRO-INDICATEURS

Distributeurs: Kynmore Engineering Co. Ltd, 19 Buckingham Gate, London, W.C.2

(Illustration à la page 53)

Le micro-indicateur Teldix, model MA-11, fabriqué par la société Teldix, une filiale allemande de Telefunken et Bendix, est maintenant vendu par la société Kynmore Engineering Co. Ltd.

Chaque micro-indicateur affiche les chiffres 0 à 9 ainsi qu'un jeu de barres de couleur mais des éléments spéciaux peuvent être fournis qui affichent des lettres, des symboles ou n'importe quels signes voulus. Les caractères en blanc sur fond noir mesurent de large × de hauteur. Ils apparaissent sur un écran frontal mat, monté sur un coffret en plastique noir, à partir de onze lampes à faisceau individuel sur chaque élément. Les dimensions hors-tout d'un micro-indicateur sont: 8,6 mm de large × 13 mm de profondeur × 46,2 mm de longueur. Des cassettes peuvent être fournies pour recevoir n'importe quel nombre d'élements montés côte à côte pour contituer une décade ou d'autres indications. Des cassettes normales en aluminium léger de couleur noire peuvent être fournis pour y monter jusqu'à six éléments.

Le micro-indicateur lui-même, c'est à dire l'affichage individuel moins la cassette, se compose de trois parties (une douille, un élément à lampe et un élément d'affichage) qui sont toutes fermement fixées. Durant l'usage, la douille est normalement vissée à l'arrière de la cassette et lélément à lampe ainsi que l'élément d'affichage sont vissés ensemble. En cas de panne de lampe, ces deux derniers éléments peuvent être démontés instantanément de la douille et retirés de l'avant de la cassette sans exiger l'emploi d'outils. L'élément d'affichage et l'élément à lampe de remplacement sont alors insérés à nouveau à l'avant de la cassette, puis enfichés dans la douille à l'aide du connecteur imprimé miniature. Les divers éléments sont à ressort afin de les maintenir fermement ensemble.

Le choix des circuits logique est laissé à l'utilisateur la plupart des types de circuits logiques à décades transistorisés convenant à cet usage. Le fonctionnement se faisant sur 1,5 V à 70 mA (±5 mA pour une durée de lampe maxima), la consommation de courant n'est que de 0,1 W. Les caractères sont visibles à des angles maximum de 30° C à partir de la normale à la lumiére du jour. La durée nominale d'une lampe est de 500 heures.

EE 77 754 *pour plus amples renseignements*

COMPTEUR DE FRÉQUENCE

Hewlett-Packard Ltd, Dallas Road, Bedford

(Illustration à la page 53)

Un nouveau compteur de fréquence à cinq chiffres d'un modèle compact, le type 5534 A, vient d'être introduit sur le marché par la société Hewlett-Packard Ltd.

Il peut effectuer la mesure moyenne de périodes simples et multiples par plots de décades allant jusqu'à 10^5 périodes, ainsi que l'intervalle de temps actionné par impulsions é la fréquence de la base de temps de 100 kHz ou à n'importe quelle fréquence appliquée extérieurement et ce jusqu'à 2 MHz. En outre, il mesure le rapport de fréquence ou les multiples de fréquence et compte les événements aléatoires.

L'affichage clair et bien défini des résultats est obtenu grâce aux tubes indicateurs numériques à cathode froide. L'emmagasinage de lectures donne un affichage continu de la mesure la plus récente, même lorsque l'appareil effectue une nouvelle mesure. Le temps d'echantillonnage est donc réduit au minimum.

La haute sensibilité de l'appareil et sa résistance d'entrée de 1 MΩ (shuntée par 15 pF) lui permettent de détecter et de mesurer des signaux jusqu'à 100 mV efficaces.

Les commandes sont simples et fonctionnelles; la décimale est placée automatiquement. L'appareil se compose exclusivement de semi-conducteurs. Le coffret peut être monté sur bâti ou banc d'essai et sa structure à plaquettes imprimées interchangeables lui donne une grande accessibilité en vue de son entretien.

EE 77 755 *pour plus amples renseignements*

ENREGISTREUR À CIRCUIT DE RETARD

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustration à la page 53)

Cet appareil a été mis au point par la Gresham Lion Electronics Ltd, pour l'enregistrement et la reproduction de signaux acoustiques ou de phénomènes transitoires sur un circuit continu de bande magnétique.

Trois longueurs de boucles peuvent être utilisées: 375 cm, 280 cm ou 187,5 cm. Dans l'enregistreur standard deux vitesses de bande sont prévues: 18,75 cm/sec et 9,37 cm/sec.

A la vitesse la plus élevée, les trois longueurs de boucles donnent des temps de recylage de 20 sec, 15 sec et 10 sec, les chiffres correspondants pour les vitesses inférieures étant de 40 sec, 30 sec et 20 sec.

D'autres vitesses de bande ainsi que d'autres temps de retard peuvent être obtenus par simple changement du moteur et du nombre de poulies folles.

Quatorze voies sont prévues sur bande de 2,5 cm, chacune d'elles comportant un amplificateur d'enregistrement transistorisé à fiches et un amplificateur de reproduction, un instrument de mesure commutable étant fourni pour contrôler les niveaux d'enregistrement et de reproduction.

Un commutateur qui élimine les circuits d'effacement et d'enregistrement permet de "fixer" sur la bande une boucle complète de signaux enregistrés.

Grâce à la précision azimutale incorporée des têtes magnétiques Gresham de reproduction et d'effacement, ces dernières sont entièrement interchangeables sans le moindre réglage mécanique.

EE 77 756 *pour plus amples renseignements*

COMPARATEUR C.A./C.C.

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(Illustration à la page 54)

Cet appareil entièrement nouveau, autonome et fonctionnant sur secteur a été conçu pour mesurer les valeurs efficaces de courant et de tension alternatifs. Sa précision est de ±0,05 % dans une gamme de fréquences de 25 Hz à 20 kHz.

Une jonction sous vide unique, dont l'erreur de conversion du courant alternatif au courant continu est inférieure à 0,01 %, est utilisée comme élément de transfert; le circuit est compensé de sorte que les pièces de remplacement peuvent être fixées sans qu'il y ait lieu d'effecteur un nouvel étalonnage.

Un des avantages les plus importants de cette nouvelle conception est que la tension ou le courant inconnus sont rapidement équilibrés en fonction d'un potentiel de courant continu réglable, qui peut être ensuite mesuré avec précision à l'aide d'un courant continu de référence donnant la valeur efficace du signal appliqué. Le courant de référence provient d'un bloc d'alimentation stabilisée Zéner ayant une variation de sortie inférieure à 0,002 % pour une fluctuation de tension secteur de 10 %.

Les seuls éléments extérieurs nécessaires sont une cellule standard et un galvanomètre à réflection approprié, l'instrument étant, quand au reste, entièrement autonome, ayant son propre circuit potentiométrique incorporé. L'appareil comporte en outre un instrument de mesure incorporé qui réduit le risque de surcharger la jonction sous vide. Cet instrument de mesure, des plus utiles, permet de déterminer la valeur approximative du signal inconnu avant de régler les commandes.

L'appareil a huit gammes de tension et de courant, couvrant des valeurs de 0,5 à 300 V et de 5 mA à 3 A respectivement.

EE 77 757 *pour plus amples renseignements*

CONTRÔLEUR DE TENSION ALTERNATIVE OU CONTINUE

Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(Illustration à la page 54)

Il s'agit ici d'un appareil de contrôle universel, compact et d'une grande souplesse d'emploi, pour mesurer la résistance ainsi que les courants et tensions continus et alternatifs.

L'instrument comprend un potentiomètre Rayleigh à quatre cadrans, un étalon de transfert thermique c.a./c.c., un diviseur, de potentiel, des shunts de courant, un galvanomètre réflecteur de spot de 50 Ω ainsi que tous les commutateurs nécessaires.

Il est particulièrement utile pour les travaux en atelier ou sur place car c'est un appareil portatif que ne présente pas l'inconvénient de devoir relier plusieurs composants comme c'est généralement le cas pour ce genre d'appareils.

Les circuits peuvent être alimentés par deux accumulateurs de 2 V ou deux blocs d'alimentation secteur de 2 V à diodes Zener, type P3/S.

Son fonctionnement est des plus simples car il ne comporte que trois commutateurs pour le choix des gammes de tension, de courant et de résistance. Ces commutateurs sont à verrouillage mécanique afin d'assurer que les gammes qui ne sont pas utilisées soient retournées à la position "arrêt".

On peut également mesurer des courants supérieurs à 5 A au moyen de résistances shunt de connexion se trouvant à l'extérieur de l'appareil.

Les circuits d'entrée comprennent un enregistreur de recherche qui facilite le choix de la gamme lorsqu'on mesure une tension ou un courant c.a. ou c.c. Pour les tensions continues allant jusqu'à 1,0999 V, la mesure s'effectue directement à l'aide du potentiomètre; pour la mesure de tensions supérieures, le diviseur de potentiel est automatiquement inclus pour le choix de la gamme voulue. La tension continue est mesurée avec le potentiomètre et les shunts de courant; les gammes sont alors changées en modifiant le sélecteur de gamme de courant.

L'étalon de transfert thermique c.a./c.c. utilisé pour la mesure du courant alternatif donne une indication efficace réelle qui est indépendante de la forme d'onde. Le dispositif de transfert utilisé consiste en une jonction thermique sous vide à erreur d'inversion réduite. Le circuit est conçu de manière à ce que la précision de la mesure soit entièrement indépendante des caractéristiques de la jonction sous vide.

Pour mesurer la résistance, l'un des côtés du potentiomètre Rayleigh est utilisé comme bras variable d'un pont de Wheatstone. Les rapports de pont sont actionnés par commutateur.

L'instrument est enfermé dans un coffret en bois de teck avec couvercle métallique. Les commutateurs sont du type "Cropico" S.P.1. Leur résistance de contact est réduite et leur isolement est élevé. Ils comprennent en outre un mécanisme de guidage positif. Des instruction d'emploi détaillées sont fournies avec chaque appareil.

Quarante gammes sont prévues, couvrant une résistance de 0 à 1 MΩ, une tension de 0 à 600 V c.c. et de 2 à 600 V c.a. (25 Hz à 10 kHz), et un courant de 0 à 5 A c.c. et de 5 mA à 5 A c.a.

EE 77 758 *pour plus amples renseignements*

MODULE D'ENTRAÎNEMENT DE THYRISTORS

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

(Illustration à la page 54)

La société Mullard Ltd vient de mettre au point un module d'entraînement de thyristors MY5000, destiné à simplifier l'emploi des thyristors dans une gamme étendue d'applications électriques ordinaires. C'est le module le plus simple de sa série et il peut entraîner en même temps jusqu'à deux thyristors à courant élevé de 70 A; en quelque sorte, comme les deux bras d'un circuit en pont monophasé.

Bien que des modules d'entraînement plus compliqués soient nécessaires pour la commande de vitesse de moteurs de grande précision avec des circuits de réaction complexes, ce module à bas prix constitue une solution économique et satisfaisante à bon nombre d'applications simples de commande de vitesse de moteur, de commande de fours de réduction de lumière ainsi que de commande de courant alternatif par dispositif à thyristors "dos-à-dos".

L'un des problèmes qui se posent aux utilisateurs de thyristors est celui de pouvoir fournir des impulsions d'entraînement sûres et qui puissent amorcer les thyristors et contrôler également la position de l'impulsion au moyen d'un élément de contrôle ou d'un circuit de réaction. Chaque application exige naturellement son propre montage à réaction, mais en utilisant un module Mullard qui fournit des impulsions d'entraînement de thyristors à tolérance équilibrée, seuls les problèmes de réaction et de courant continu subsistent. Les problèmes les plus communs seront l'objet de rapports techniques Mullard.

Le module de déclenchement MY5000 est totalement protégé contre les effets ambiants par son revêtement en matière plastique qui protège les composants de circuit et les interconnexions contre tout endommagement dû aux vibrations ou au choc mécanique.

EE 77 759 *pour plus amples renseignements*

RÉSERVOIRS DE LIGNES À RETARD

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Illustration à la page 54)

L'emmagasinage en série de données comportant jusqu'à 1 000 chiffres binaires peut être assuré de façon économique à l'aide d'une nouvelle gamme de lignes à retard à ultrasons dont toutes les pièces sont normalisées.

Désignées sous le nom de "modules à retard YL2108", chacun des éléments de la gamme est fourni complet avec des circuits d'entrée et de sortie constitués de corps solides et de composants magnétostrictifs à action différée par fil. N'importe quelle action dfférée peut être obtenue entre 100 μsec et 3,2 msec, chaque type d'élément étant réglable dans une gamme de 20 μsec.

Les éléments peuvent être utilisés dans une gamme de températures de —10° C à +60° C avec des vitesses de chiffres binaires maxima de 350 kHz. Un bloc d'alimentation de 12 V est nécessaire, ce dernier pouvant être positif, négatif ou symétrique. Ces éléments sont donc indiqués pour la plupart des types de systèmes logiques à transistors p.n.p. et n.p.n., et ils sont entièrement compatibles avec les blocs de circuits numériques de la série M.E.L.I.

Les lignes à retard fonctionnent normalement suivant le mode cyclique, qui amplifie la sortie et la renvoie à l'entrée, l'information étant retenue à condition que l'alimentation soit maintenue. L'information peut être extraite à n'importe quel moment, avec un temps d'accès maximum égal au retard total dans la ligne.

De nombreuses utilisations sont envisagées pour les modèles les plus simples de matériels de traitement des données. Les lignes peuvent servir de mémoires temporaires dans les petites machines à calculer, les machines à facturer, les calculatrices "sur ligne" ainsi que dans des appareils analogues. Elles peuvent être utilisées également en liaison avec des mémoires permanentes telles que les réservoirs à noyau. Les lignes peuvent garder les informations pendant de courtes périodes de temps permettant d'effectuer les opérations de calcul ou de séquence.

Les lignes sont enfermées dans des boîtiers en acier bordé mesurant 23,49 cm × 17,14 cm × 3,81 cm. Ces boîtiers peuvent être montés dans n'importe quelle position. Ils peuvent être montés également sur le système standard de bâtis M.E.L. Un des côtés étroits de l'élément constitue le panneau frontal comportant un point de contrôle de sortie et une commande de retard. Les bornes d'entrée et de sortie de puissance et de signaux se trouvent à l'arrière.

EE 77 760 *pour plus amples renseignements*

CONTRÔLEUR DE BRUIT BLANC

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustration à la page 55)

La société Marconi Instruments Ltd a

réalisé un nouveau contrôleur de bruit blanc, type OA 2090, qui est entièrement transistorisé et conçu pour le contrôle de liaisons hertziennes multivoies et par câbles dont la capacité va jusqu'à 2700 voies. Ce contrôleur est conforme aux recommandations du C.C.I.R. et comprend deux éléments. Son poids et ses dimensions sont considérablement plus réduits que ceux des modèles antérieurs.

Dans les systèmes multivoies, il est essentiel de réduire au minimum le brouillage d'intermodulation dans les voies. Cette diaphonie est produite principalement par la distorsion non-linéaire et des phases et elle est audible sous la forme d'une interférence ressemblant à des bruits erratiques qu'on détermine en mesurant le rapport de la puissance de bruit.

Le contrôleur fonctionne suivant les principes suivants: le bruit blanc du niveau de puissance et de la largeur de bande appropriés est appliqué au système, simulant très fidèlement un système téléphonique multivoies à pleine charge. Un filtre à bande étroite de fréquences non transmises est posé entre la source de bruit blanc et le système, produisant ainsi une voie silencieuse. Un récepteur accordé à la voie silencieuse est utilisé pour comparer le niveau de bruit produit par l'intermodulation des composants du bruit blanc occupant le reste de la bande de fréquences avec le signal original dans la voie.

Le générateur de bruit utilise une diode semi-conductrice spéciale comme source de bruit. Il peut contenir jusqu'à 9 filtres à changement rapide: tant les filtres à limitation de bande que ceux à élimination de bande sont incorporés. Les éléments à filtre comprennent des commutateurs avec des clavettes à bras articulé à code de couleurs et les positions dans le circuit de tous les filtres sont occupées avant que n'intervienne le contrôleur de puissance; les limites de bande de bruit et les positions des fentes sont donc choisies par simple commutation, cependant que la référence de niveau de puissance est contrôlée de façon continue. La charge du système est réglée par un atténuateur à la sortie.

Le récepteur peut comporter jusqu'à six fréquences ponctuelles de réception et il comprend également des filtres interchangeables qui correspondent aux fréquences ponctuelles du générateur. L'atténuateur d'entrée est marqué de manière à ce que le rapport de puissance de bruit en décibels puisse être lu directement.

Les dimensions des éléments sont comme suit: 19,05 cm × 46,99 cm × 43,18 cm. Le poids est de 11,75 kg. Il existe également des versions pour montage sur bâti.

EE 77 761 *pour plus amples renseignements*

NOYAUX DE FERRITE

Standard Telephones & Cables Ltd, Magnetic Materials Division, Edinburgh Way, Harlow, Essex

(Illustration à la page 55)

La division des matériaux magnétiques STC vient d'ajouter à sa gamme de noyaux magnétiques à haute performance une nouvelle série de noyaux conformes aux normes internationales. Ce nouveau type de noyau se caractérise par le fait que son mécanisme de réglage est entièrement contenu à l'interieur du noyau. Ainsi les assemblages de noyau occupent l'espace minimum dans le matériel et demeurent stables même dans les conditions les plus défavorables.

Il existe trois formats de noyaux, en deux matériaux, à savoir les formats de 26 mm, 18 mm et 14 mm, dans les matériaux SA501 et SA503. D'autres formats en divers matériaux seront ajoutés progressivement de manière à former une gamme complète pour toutes les applications.

Les matériaux SA501 et SA503 sont tous deux à stabilité élevée et à faible perte, pouvant être utilisés dans les circuits accordés jusqu'à 500 kHz et dans les circuits non accordés jusqu'à 1 MHz. Le matériau SA501 a un facteur de température très réduit et il peut être utilisé avec les condensateurs au mica. Le matériau SA503 a un facteur de température positive contrôlée et il peut être utilisé avec les condensateurs au polystyrène. Les noyaux IEC conviennent parfaitement au montage sur châssis et sur plaquette de circuit imprimé. La société Standard Telephones & Cables fournit des bobines Delrin pour tous les formats.

La société STC effectue un contrôle rigoureux de la composition chimique et de la réactivité, par une vérification continue durant la fabrication des noyaux IEC. Cette vérification porte en particulier sur la microstructure des ferrites et ce procédé de contrôle assure la fabrication de noyaux d'une qualité inégalée.

EE 77 762 *pour plus amples renseignements*

INDICATEUR HORAIRE

Venner Ltd, Kingston By-Pass, New Malden Surrey

(Illustration à la page 55)

La société Venner Ltd, a créé un indicateur horaire synchrone actionné par moteur, type SHR1, pouvant être utilisé avec n'importe quel appareil fonctionnant sur fréquence alternative contrôlée. L'enregistrement s'effectue par fractions d'un dixième d'heure jusqu'à 99 999,9 heures, avec réenclenchement automatique à zéro.

Les applications du nouvel indicateur comprennent les essais de durée, le contrôle de consommation des matériaux, l'enregistrement de la durée de fonctionnement de pompes et de matériel électronique, ainsi que l'indication des périodes d'entretien de tours automatiques et de matériel d'usine analogue.

L'indicateur est livrable en deux modèles: l'un prévoyant le montage de surface avec entrée de câble ou de conducteur, l'autre prévoyant le montage noyé. Une plaquette de connexions à quatre directions permet d'utiliser l'indicateur à partir de trois tensions alternatives: 100 à 120 V, 200 à 250 V, 380 à 460 V, 50 Hz. La consommation électrique est de 2,2 W à 240 V. Les dimensions du modèle à montage noyé sont de 7,62 cm de largeur, 6,99 cm de profondeur et 5,08 cm de hauteur.

L'indicateur peut être fourni sur commande pour l'enregistrement des minutes plutôt que des heures.

EE 77 763 *pour plus amples renseignements*

THERMOSONDE

Distributeurs: The Solartron Electronic Group Ltd, Victoria Road, Farnborough, Hampshire

(Illustration à la page 56)

La thermosonde est un instrument thermoélectrique portatif, conçu pour provoquer des variations de température dans les petits composants. Deux éléments de refroidissement thermoélectriques sont montés dans une sonde munie d'assemblages à pointe amovibles. Des commandes simples et un grand indicateur de température sont montés sur le panneau frontal d'un solide coffret métallique contenant un bloc d'alimentation en courant continu et un système échangeur de chaleur.

L'instrument peut élever ou abaisser la température de composants individuels en circuit afin de faciliter la conception et la mise au point de prototypes. Des pointes de sonde de différentes forme sont prévues pour assurer le contact thermique maximum avec le composant soumis à l'essai. D'autres applications possibles comprennent le contrôle de marchandises entrant en magasin, la recherche des dérangements ou des pannes et le frettage de petites pièces par chauffage ou refroidissement directs. L'appareil trouvera certainement aussi des applications dans les laboratoires de recherche médicale et chimique.

La gamme de température de la thermosonde s'étend de -20° C à +70° C et le temps de réponse est de deux minutes pour atteindre 60 % de la température finale.

La thermosonde modèle TP-10 utilise des éléments de refroidissement thermoélectrique pour pomper de la chaleur entre la zone de contrôle et l'atmosphère ambiante. Dans le mode à refroidissement, la chaleur est extraite par des pointes spéciales d'un objet, tel qu'un composant électronique dans un circuit actif, et transmise par conduction aux surfaces des éléments thermoélectriques. De là, la chaleur est transmise aux faces opposées des éléments par le courant continu appliqué, puis conduite vers le liquide de refroidissement. Le liquide circule alors à travers l'échangeur de chaleur liquide/air et la chaleur est transmise à l'atmosphère. Dans le mode de chauffage, le flux de chaleur est inversé lorsqu'on change la direction du courant continu appliqué. La chaleur, tirée de l'atmosphère, est pompée dans la zone d'essai. La quantité de chaleur pompée est en fonction du niveau de courant continu. La température est "détectée" par un thermocouple cuivre/constantan relié à un pyromillivoltmètre Weston.

L'instrument est fabriqué par la société Daystrom Ltd du Canada et il est vendu dans le Royaume-Uni par The Solartron Electronic Group Ltd.

EE 77 764 *pour plus amples renseignements*

TRANSDUCTEUR MAGNÉTIQUE MINIATURE

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(Illustration à la page 56)

La société Advance Controls Ltd vient d'ajouter un nouveau transducteur magnétique miniature, le modèle MT3SR, à sa gamme actuelle. Ce transducteur ne contient aucune partie mobile et produit des impulsions à partir d'arbres rotatifs et de roues d'engrenages sans contact physique aucun. Ces impulsions peuvent être utilisées pour actionner des contrôleurs de rotation, des commutateurs de vitesse, des tachymètres analogiques ou numériques, des instruments de mesure du rapport, des indicateurs de différence de vitesse et de nombreuses autres applications de contrôle.

Le transducteur est monté avec sa pièce polaire près du parcours des dents d'engrenage mobiles et avec l'espacement recommandé de 0,03 à 0,06 pouces (laissant ainsi un peu de jeu aux coussinets). Les impulsions de plus de 300 mV d'amplitude sont produites avec des vitesses périphérales minima de 5 cm/sec.

Le nouveau transducteur est hermétiquement scellé dans un robuste coffret en laiton convenant pour presque tous les usages industriels. Il mesure 5 cm de long et son filetage est de 5/8 B.S.F., pour en faciliter le montage. Douze autres modèles courants pour diverses applications sont offerts à la vente et le nombre de modèles spéciaux pouvant être fournis s'élève à 60.

EE 77 765 *pour plus amples renseignements*

COMMUTATEUR DE GUIDE D'ONDES

The Marconi Co. Ltd, Chelmsford, Essex

(Illustration à la page 56)

Un nouveau commutateur de guide d'ondes entièrement électronique avec une puissance de 2 W, couvrant la gamme de fréquences de 7,75 à 8,5 VHz peut être obtenu maintenant de la Marconi Co. Ltd. L'atténuation du commutateur est de 50 dB, et il a été conçu pour pouvoir être utilisé avec le matériel de communications microondes dont il améliorera considérablement la sûreté de fonctionnement, la vitesse, l'économie d'espace et l'efficacité par rapport aux méthodes de commutation mécanique employées jusqu'ici.

Le nouveau commutateur est du format de guide d'ondes No. 15 et consiste en une longueur de guide d'ondes spécialement fabriquée contenant une pièce en ferrite avec une bobine enroulée autour du corps extérieur du guide d'ondes pour produire un champ magnétique longitudinal. Lorsque ce champ est excité l'atténuation s'élève rapidement à 50dB. Le temps de commutation peut être réalisé en moins d'une milliseconde par un petit circuit de commutation transistorisé.

EE 77 766 *pour plus amples renseignements*

BLOC D'ALIMENTATION À DIODES TUNNEL

Coutant Electronics Ltd, 3 Trafford Road, Richfield Estate, Reading, Berkshire

(Illustration à la page 56)

Les tensions relativement faibles des circuits à diodes tunnel exigent des methodes spéciales, en particulier lorsque les courants sont de plusieurs ampères.

La société Coutant Electronics Ltd a conçu un bloc d'alimentation stabilisée s'adaptant aux lignes de bandes imprimées normales. Les connexions de sortie s'effectuent en soudant les lignes de bande directement au circuit imprimé à deux côtés se trouvant dans le bloc d'alimentation.

La tension de sortie est à variation continue entre 0 et 100 mV à un courant de 8 A et deux alimentations auxiliaires de +5 V et −5 V à 6 A sont fournies pour les circuits à transistors connexes.

Les signaux d'erreurs éloignés sont transmis de la charge à l'amplificateur de commande à travers des câbles et connexions coaxiaux miniature. Le temps de réponse du stabilisateur au changement de degré du courant de charge est voisin du minimum théorique pouvant être atteint et dépend principalement de la longueur de la ligne de bande employée entre le bloc d'alimentation et le point de perception.

Le bloc d'alimentation utilise un montage de semi-conducteurs au silicium et il peut fonctionner dans des températures ambiantes s'élevant à un maximum de 60° C. Il a été conçu pour montage sur bâti standard de 48,26 cm et il occupe une hauteur de panneau de 13,33 cm.

EE 77 767 *pour plus amples renseignements*

FRÉQUENCEMÈTRE NUMÉRIQUE

Dawe Instruments Ltd, Western Avenue Acton, London, W.3

(Illustration à la page 56)

Pour répondre à la demande d'un fréquencemètre numérique à prix modéré et convenant à une gamme étendue d'applications industrielles générales et de laboratoire, la société Dawe Instruments a réalisé le fréquencemètre numérique, type 719A. Comprenant un cadran indicateur illuminé à 5 chiffres et un oscillateur piézo-électrique de 10 kHz, ce fréquencemètre est indiqué pour toute une série d'utilisations comprenant la mesure de fréquence, de périodes et de vitesse (en tours/minute), ainsi que la comptage des impulsions aléatoires.

Le fréquencemètre type 719 A est basé sur les principes classiques du compteur-minuterie numérique, la forme d'onde d'entrée électrique étant amplifiée et limitée de manière à produire un train d'impulsions. Ces dernières sont comptées pendant une période de déclenchement déterminée ou, inversement, elles sont elles-mêmes utilisées pour contrôler le déclenchement. Pour la mesure de fréquence, des impulsions à mouvement d'horlogerie émanant de l'oscillateur à cristal sont utilisées en liaison avec des diviseurs de décades pour ouvrir et fermer le dispositif de déclenchement pendant une période préréglée de 0,01 à 10 sec, au cours de laquelle les impulsions d'entrée sont comptées et affichées. La vitesse est mesurée de la même manière, pendant une période de 6 sec, l'indication directe étant fournie en tours/minute. Pour la mesure des périodes, la procédure est inversée, le nombre des impulsions de l'oscillateur piloté au quartz (ou une division de décades de ce nombre) étant compté pendant un cycle de l'entrée. L'instrument peut être utilisé également pour le travail "chronométrique", en comptant le nombre d'impulsions durant une période d'entrée fixe. Les impulsions aléatoires peuvent aussi être comptées, à condition que l'intervalle entre les impulsions successives ne soit pas inférieur à 0,1 μsec.

La fréquence peut être mesurée du courant continu à 10 MHz, les périodes et les impulsions aléatoires pouvant être mesurées jusqu'à 99999 sec. La précision totale est de ±1 comptage, ± la précision du quartz, ce dernier étant réglé à une partie dans 10^6 à la température de la chambre. L'affichage numérique indique la décimale et le "dépassement de comptage", l'indication finale étant visible pour une période préréglée de 1 sec à 15 sec ou jusqu'au réenclenchement manuel. Le fréquencemètre est entièrement transistorisé et son circuit a été considérablement simplifié par l'emploi d'un nouveau diviseur de décades (brevet demandé).

Le fréquencemètre type 719 A convient à de nombreuses utilisations comprenant la tachymétrie, l'étalonnage de fréquence, le minutage de contacts de relais, les travaux "chronométriques" synchronisés et la mesure de couple et de phase. Il est particulièrement indiqué pour les utilisations où un fréquencemètre numérique était considéré jusqu'ici comme trop coûteux.

La consommation électrique de l'appareil n'est que de 15 W (secteur) et son poids est de 15 kg. Il mesure 52 cm × 27,9 cm × 30,4 cm de profondeur, le coffret pouvant être monté sur bâti ou sur banc d'essai.

EE 77 768 *pour plus amples renseignements*

AMPLIFICATEUR OPÉRATIONNEL

Bendix Electronics Ltd, High Church Street, New Basford, Nottingham

(Illustration à la page 57)

La société Bendix Electronics Ltd. vient de mettre au point un nouvel amplificateur stabilisé à relais modulateur et constitué de corps solides qui porte la désignation 470. L'application d'une réaction autour de l'amplificateur produit une caractéristique de gain précise qui est entièrement en fonction des impédances de réaction. De plus, l'entrée est pratiquement au potentiel nul et aucun courant ne passe à l'amplificateur, ce qui représente une fonction de sommation idéale pour les courants d'entrée. L'amplificateur 470 utilise des transistors planaires épitaxiaux pour assurer le maximum de fiabilité; sa gamme de températures s'étend de 0 à 70° C et son gain est de 8×10^6 au courant continu, se réduisant à l'unité à 2 MHz.

Le gain de l'amplificateur auxiliaire est de 2 000, se réduisant à 20dB par décade à partir de 0,045 Hz. Le gain de l'amplificateur à large bande est de 4 000 et tombe à 20 dB par décade à partir de 1 kHz. L'impédance d'entrée est de 6 kΩ et la sortie de 10 V dans une charge de 1 kΩ.

EE 77 769 *pour plus amples renseignements*

MINUTERIE DE PROCESSUS

Distributeurs: Rayleigh Instruments Ltd, 271 Kiln Road, Thundersley, Essex

(Illustration à la page 57)

Une nouvelle gamme de minuteries de haute précision et de relais à action différée fabriqués par la société DOLD & Söhne d'Allemagne Fédérale vient d'être introduite au Royaume-Uni par la société Rayleigh Instruments Ltd.

Ces instruments du type à moteur synchrone sont de construction éprouvée et conformes aux spécifications électriques et de minutage les plus rigoureuses. Ils sont livrables en différents modèles, dont certains peuvent recevoir jusqu'à 5 gammes de temps différentes. Sur les instruments multigammes, le choix de la gamme voulue se fait au moyen d'une simple vis de commande. Les périodes de retard peuvent être fixés en n'importe quel point entre les limites de 0,1 sec à 120 H, selon la gamme de l'instrument choisi.

Ces instruments peuvent être munis sur demande de contacts fonctionnant soit au début soit à la fin du cycle de temps.

EE 77 770 *pour plus amples renseignements*

CONTRÔLEUR DE PUISSANCE AUTOMATIQUE

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustration à la page 57)

La société Kent Precision Electronics Ltd produit maintenant une gamme de contrôleurs automatiques constitués de corps solides d'un modèle très perfectionné et prévus pour l'utilisation avec les régulateurs motorisés. La commande automatique normale est effectuée en commutant les régulateurs motorisés de manière à élever ou réduire la sortie, selon les besoins.

La gamme des instruments KPE fonctionne avec un instrument de contrôle étalonné standard constitué de corps solides pour le contrôle proportionnel d'un moteur commutateur à courant continu. Pour les erreurs importantes, le moteur est commuté intégralement et, à l'intérieur d'une bande proportionnelle d'environ 2 %, il ralentit pour atteindre le point fixé sans surmodulation ni recherche d'une ligne libre. Un espace mort de moins de 0,5 % est facilement réalisé dans toutes les applications normales. Lorsque le contrôle arrêt-marche ordinaire est exigé pour la gamme commerciale courante des régulateurs motorisés, un instrument KPE standard peut être fourni à cet effet. Cet instrument utilise un appareil de contrôle étalonné constitué de corps solides qui actionne un relais standard pour le contrôle arrêt-marche.

Le fonctionnement suivant un programme, comme par exemple dans le cas où le contrôle constant d'une densité de courant est exigé pour les réservoirs de galvanoplastie, est effectué en introduisant des entrées opérationnelles secondaires dans l'instrument de contrôle étalonné standard constitué de corps solides. On voit dans notre illustration un contrôleur standard de densité de courant constant.

Ces instruments robustes sont fournis pour montage sur panneau ou sur baies, et comme ils n'ont pas de parties mobiles, les systèmes de contrôle automatique utilisant la série d'instruments KPE sont assurés d'un fonctionnement sûr et sans nécessité d'entretien.

EE 77 771 *pour plus amples renseignements*

ATTÉNUATEUR AUTOMATIQUE

Sangamo Weston Ltd, Great Cambridge Road, Enfield, Middlesex

La société Sangamo International produit maintenant un atténuateur automatique, type SA-26. Ce dispositif répond à la demande des utilisateurs d'appareils de traitement de données analogiques pour la plupart des cas où des pointes occasionelles ou des amplitudes imprévisibles surentraînent le système indicateur ou enregistreur, qu'il s'agisse d'un enregistreur à bande magnétique, d'un oscillographe, d'un enregistreur à diagrammes ou d'un appareil de report sur carte.

Le modèle SA-2L atténue des signaux au-dessus d'un niveau de déclenchement réglable, dans une proportion déterminée avec précision; il étend ainsi la gamme dynamique de n'importe quel système à bruit constant. L'atténuateur n'est pas limité aux applications comportant les pointes. Il peut être utilisé également comme extenseur de gamme d'une précision totale sur 2 points de sélection.

EE 77 772 *pour plus amples renseignements*

Résumés des Principaux Articles

Un analyseur de spectres à bande X et un générateur de balayage par R. W. Eldred

Résumé de l'article aux pages 2 à 6

Cet article décrit brièvement un analyseur de spectres à bande X et un générateur de balayage combinés. Le générateur de balayage dérive de la méthode particulière utilisée pour assurer une sensibilité constante à l'analyseur de spectre dans la gamme de balayage électronique. L'article décrit également une méthode permettant d'utiliser un ondemètre interne du type à absorption dans un circuit à réglage automatique de fréquence, pour stabiliser la fréquence de l'oscillateur local dans des conditions de "balayage manuel".

Expériences effectuées dans la production de puissance MHD par G. J. Womack

Résumé de l'article aux pages 7 à 14

Cet article traite des résultats obtenus à l'aide d'un petit générateur MHD entraîné par combustion d'oxygène-propane gazeux. Un aimant permanent relativement grand fut utilisé pour produire une densité à flux constant de 9 800 gauss. Les produits de la combustion furent ensemencés avec environ 1,6% de potassium. Les expériences furent effectuées tant avec des électrodes à carbone chaud qu'avec des électrodes de cuivre refroidies à l'eau.

La conductivité électrique du fluide de travail a été déterminée à l'aide des deux types d'électrode en mesurant les caractéristiques I-V dans des conditions de champ électrique appliqué et induit. Pour faciliter l'interprétation de ces résultats on construisit un appareil à inversion de ligne au sodium à double faisceau qui permit de mesurer la température du gaz. On obtint un bon accord entre la conductivité électronique mesurée et la conductivité calculée théoriquement, tant pour les électrodes chaudes que pour les électrodes froides. Pour les électrodes froides le mode de conductivité employé fut électronique et non ionique, la cathode froide devant donc émettre des électrons. Le mécanisme exact de cette émission n'est pas entièrement compris, mais il est probablement dû au

dépot de carbone sur la surface des électrodes. Ce carbone est probablement amorphe et d'une porosité élevée et d'une faible conductivité thermique. Cette couche approche donc les conditions de stagnation du courant gazeux qui serait suffisant pour produire une émission thermionique élevée.

Les mesures de conductivité électriques ont permit de déterminer les coupes transversales de collision, ces dernière étant ensuite comparées aux valeurs prédites théoriquement et calculées par la méthode de Frost. Un accord raisonablement satisfaisant fut obtenu.

Un bruit de tension à basse fréquence fut observé.

Des mesures de générateur Hall furent effectuées pour déterminer $w_e t_e$.

Un transformateur d'entrée à faibles fuites de courant à la terre par D. G. Wyatt

Résumé de l'article aux pages 16 à 19

Les sources d'erreurs produites par les transformateurs d'entrée utilisés dans les circuits de mesure électrique sont étudiées dans cet article qui indique en détail la construction d'un transformateur à faible fuite de courant à la terre. Il décrit en particulier les circuits de neutralisation employés. Le comportement d'un transformateur complet est également décrit.

Une calculatrice de moyenne à bon marché pour l'analyse "sur ligne" de données neurophysiologiques par D. Burns, W. Ferch et G. Mandl

Résumé de l'article aux pages 20 à 24

En raison de l'intérêt croissant porté au comportement de moyenne-temps des cellules nerveuses, il s'est produit une demande accrue d'instruments pouvant fournir rapidement une classification "sur ligne" d'intervalles de temps. Cet article décrit donc une calculatrice spéciale à un prix relativement réduit, pouvant effecteur des analyses d'intervalles, ainsi que la corrélation numérique, transversale ou automatique.

Multivibrateurs de déphasage nul par A. Pugh

Résumé de l'article aux pages 24 à 25

Les recherches effectuées dans le domaine des circuits d'impulsions positives et négatives ont conduit à la réalisation de circuits multivibrateurs à transistors comportant des étages à base à la masse, par rapport aux étages à émetteur à la masse. Cet article signale les propriétés plutôt originales d'une série de circuits conçus selon le principe précité et indique quelques applications.

Filtres passe-bande basse fréquence à haute stabilité par P. G. Simpson

Résumé de l'article aux pages 26 à 31

Cet article traite de filtres passe-bande basse fréquence d'une stabilité très élevée dont la stabilité des phases est supérieure à une partie dans 10^6. Les aspects théoriques du choix de circuits accouplés sont examinés et quelques courbes à angle de phase et à atténuation précis sont indiquées. Un bref aperçu est donné de leur dérivation. Enfin, quelques caractéristiques pratiques de réalisation des filtres complets sont mentionées.

Deux limites distinctes pour oscillateurs à transistors à réaction par B. Shahzadi

Résumé de l'article aux pages 32 à 34

Deux critères différents pour l'oscillation soutenue peuvent être trouvés dans la documentation sur les oscillateurs à transistors à réaction. Les auteurs de cette documentation indiquent habituellement l'un de ces critères et ignorent l'autre par inadvertance.

Dans cet article, ces critères sont calculés et examinés pour plusieurs types d'oscillateurs à réaction.

Un indicateur combiné de tension et de terre pour alimentation à courant continu par G. T. Edwards

Résumé de l'article aux pages 34 à 35

Il s'agit ici d'un circuit simple pouvant être utilisé dans les blocs d'alimentation à courant continu. Il indique la sortie ainsi que celles des bornes de sortie qui est à la terre. Il peut être employé, entre autres, lorsqu'un bloc alimente un certain nombre de sorties éloignées.

Prérégulateur de redresseur piloté au silicium pour alimentation de puissance transistorisée par D. Ophir

Résumé de l'article aux pages 36 à 40

L'auteur décrit un prérégulateur de redresseur piloté au silicium d'une grande fiabilité utilisé pour un bloc d'alimentation transistorisé. Le prérégulateur est mis en action à une fréquence élevée à partir d'une source de courant continu.

Compteur de décades réversible avec réenclenchment logique à courant continu par J. H. Smith

Résumé de l'article aux pages 41 à 43

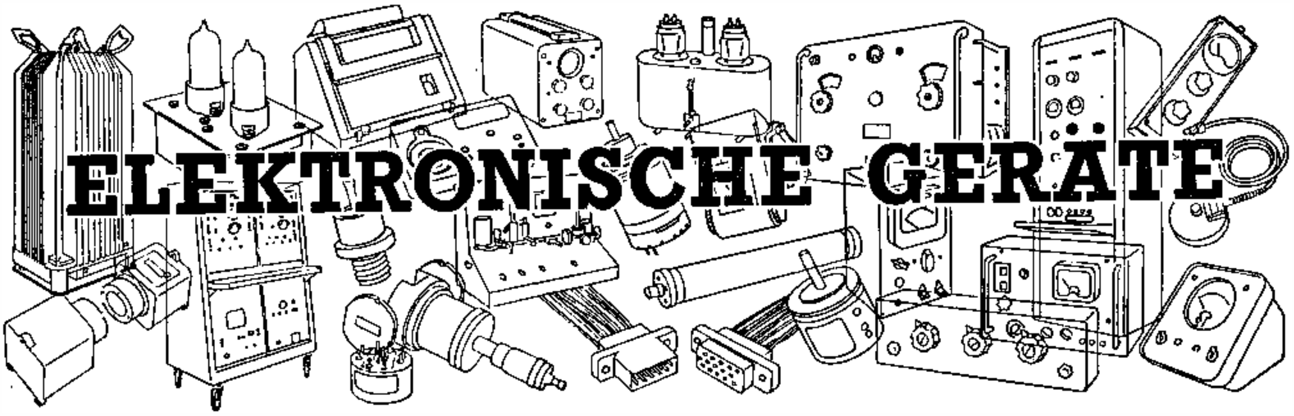
L'auteur décrit un compteur de décades réversible d'une grande fiabilité avec réenclenchement forcé à courant continu. Ce compteur est prévu pour donner le même code de sortie dans les deux directions de comptage. Il décrit également la réalisation d'un compteur de décades analogue dont la sortie décimale pour les deux directions de comptage est dans le code binaire 8 421. Ces compteurs ne sont pas auto-complémentaires.

Le fonctionnement d'un multivibrateur à couplage magnétique avec noyaux de circuit BH carrés par G. Smiljanic

Résumé de l'article aux pages 44 à 46

L'auteur étudie quelques unes des caractéristiques d'un multivibrateur à couplage magnétique et à réglage de fréquence pour l'entraînement de noyaux de transformateurs à cadres d'hystérésis carrés. Il montre que la fréquence du multivibrateur dépend linéairement de la tension d'entrée. La fréquence peut être réglée en changeant les résistances. Elle peut être stabilisée par des diodes Zener contre les variations de tension d'entrée ou de température. Enfin elle peut être variée dans une gamme étendue par un courant continu de contrôle maintenant la forme d'onde carrée des impulsions de sortie à une amplitude constante.

ELEKTRONISCHE GERÄTE



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 52 bis 57

Digitalmagnetband-Laufwerk

Plessey-UK Ltd, Control & Nucleonics Division, West Leigh, Havant, Hampshire

(Abbildung Seite 52)

Der Bandantrieb des M3000 arbeitet nach einem neuen Luftdruckprinzip, das die bei Laufwerken mit Andruckrollentrieb auftretenden Nachteile beseitigt. Mit einem Vakuum-Sammelbehälter und kontinuierlichen Spulenservo hat der M3000 sehr hohe Grundleistungsdaten bei hoher Zuverlässigkeit und niedrigen Kosten, die den Anforderungen der Konstrukteure und Benutzer von Elektronenrechnern entsprechen. Auf Wunsch ist auch eine Anzahl anderer Ausführungen zu weniger anspruchsvollen Pflichtenblättern lieferbar.

Eine gleichmässige Verteilung der Beschleunigungskräfte über die Bandbreite gibt geringen Bandschräglauf, ein Kennzeichen des pneumatischen Bandantriebes. Elektronische Kompensation für Bandschräglauf ist nicht erforderlich.

Sanftes Handhaben und Aufspulen des Bandes ist ein Kennzeichen des M3000, und die Gesamtbandabnutzung ist sehr gering. Das ist vor allem auf die über eine grosse Fläche der Triebrolle angreifenden niedrigen Antrieb-Beschleunigungskräfte und die Tatsache, dass die Triebrolle nicht auf das Band aufprallt, zurückzuführen. Schnelle und gleichmässige Bandbeschleunigung reduziert die Spannungen im Band auf ein Minimum.

Für hohe Ausnutzung von Digitalrechnern ist schnelles Auswechseln der Bandspulen besonders wichtig, eine Forderung, die das geradlinige Einlegen des Bandes und die hohe Rücklaufgeschwindigkeit des M3000 erfüllen.

Das Modell M3000 wurde für hohe Zuverlässigkeit konstruiert, so dass der Benutzer nur mit geringer Wartung und Ausfallzeit zu rechnen braucht. Die einfache Konstruktion des Luftdrucktreibsystems ermöglicht die Erfüllung dieser Entwicklungsziele.

Im M3000 wird durchgehend Festkörper-Elektronik zusammen mit den neusten Methoden zur Erzielung konstanten Funktionierens über einen Bereich von Netzspannungen, Umgebungstemperaturen und Feuchtigkeit angewendet. Es ist für Standard-Einphasennetzanschluss ausgelegt; alle anderen für die Steuerung und Signalelektronik erforderlichen Spannungen werden intern erzeugt.

EE 77 751 *für weitere Einzelheiten*

Beschleunigungsmesser

Vertrieb: B. & K. Laboratories Ltd. Tilney Street, Park Lane, London, W.1

(Abbildung Seite 52)

Die neuen hochempfindlichen Bruel & Kjaer-Beschleunigungsmesser wurden für Einsatz in schwierigen Umgebungsbedingungen und bei hohen Temperaturen entwickelt.

Diese Beschleunigungsmesser haben kleine Abmessungen, geringes Gewicht und sind sehr robust konstruiert. Sie sind für die meisten Schwingungsmessungen geeignet, vor allem für Einsatz unter harten Umgebungsbedingungen wie z.B. Feuchtigkeit, hohe Temperaturen, Magnetfelder und korrodierende Atmosphäre. Sie sind in dichter und wasserundurchlässiger Bauform ausgeführt.

Vier verschiedene Typen sind lieferbar, zwei mit Wasserkühlung und Kopfanschlüssen, die anderen beiden mit Seitenanschlüssen. Eine Auswahl von Frequenzen ist ebenfalls vorhanden: die kleineren Typen arbeiten mit Titaniumbasis in Frequenzbereichen 2...8 000 Hz (2 Prozent) oder 2...12 000 Hz (10 Prozent) und die grösseren mit Stahlbasis und Frequenzbereichen von 2... 10 000 Hz (2 Prozent) oder 2...18 000 Hz (10 Prozent).

Die Empfindlichkeit der kleineren Modelle ist 10...20 mV/g, die der grösseren 40...60 mV/g, und jeder Beschleunigungsmesser wird mit seiner eigenen Eichkurve geliefert.

Alle Modelle haben niedrige Querempfindlichkeit, einen vernachlässigbaren Befestigungsfehler und können in Temperaturen bis zu 260° C arbeiten. Die wassergekühlten Modelle können in Temperaturen bis zu 1000° C betrieben werden.

Bruel & Kjaer-Beschleunigungsmesser werden einzeln mit allem Zubehör in Kästen oder in Paketen von fünf mit Verbindungskabel geliefert.

EE 77 752 *für weitere Einzelheiten*

Ausbildungs-Servosystem

Servomex Controls Ltd, Crowborough, Sussex

(Abbildung Seite 52)

Das Ausbildungs-Servosystem SS 132 wurde entsprechend den Anforderungen an eine Ausrüstung für die Ausbildung in den Grundprinzipien der Regeltechnik entwickelt, die einerseits verhältnismässig billig ist, andererseits so weit wie möglich den Anlagen entspricht, die der Student später in der Industrie antrifft. Für die Ausbildung mit hochwertigen militärischen Servosystemen sind bereits andere Ausrüstungen vorhanden.

Zur Erfüllung dieser Forderungen hat Servomex Controls Ltd das Ausbildungs-Servosystem SS 132 entwickelt—ein Positioniersteuerrungsservo, das durchgehend mit 50 Hz als Träger arbeitet. Es ist mit Modulator und Demodulator ausgerüstet und kann von einem einfachen NF-Oszillator getrieben werden; die Ergebnisse werden auf einem Oszillografenschirm oder Registriergerät dargestellt. Die Bedienelemente sind sorgfältig geeicht, so dass der Student quantitative Versuche durchführen kann, die die Bedeutung der einfachen linearen Theorie in ausreichender Weise bestätigen.

Die Anlage hat deutlich begrenzte Blöcke mit bekannten Übertragungsfunktionen und Addiereinheiten, die nur addieren. Das ist eine wesentliche Hilfe für den Studenten, da er sich nunmehr auf das Studium des Servos selbst konzentrieren kann. Eine bemerkenswerte Eigenschaft ist, dass alle Bausteine—wie z.B. Summier- und Phasenschiebereinheiten—ohne gegenseitige Einwirkung arbeiten. Der Student ist somit in der Lage, festzustellen, welches Bauelement für einen bestimmten Arbeitsablauf verantwortlich ist. Unbeabsichtigte Sättigung wird durch einen Überlastungsanzeiger vermieden, der durch eine blinkende Lampe auf Verstärkerüberlastung aufmerksam macht. Geschwindigkeitsrückführung kann zur Beseitigung viskoser Reibung und allgemeinen Erleichterung des Messens eingeschaltet werden.

Der Motor treibt die Last über eine Kupplung mit ausgeprägtem Federungswiderstand; eine grosse Schleppscheibe, die an jeder Seite der nachgiebigen Kupplung befestigt oder vollkommen entfernt werden kann, ist vorhanden. Ausserdem gibt es eine einfache Trommel für Montage auf die Ausgangswelle zwecks Anhängen bekannter Gewichte. Auf diese Weise kann der Student den Widerstand des Servos gegenüber störenden Drehmomenten messen und denselben zur Verstärkung des Systems in Beziehung bringen.

Eine akademisch gebildete Lehrkraft für die Regeltechnik hat ein Handbuch geschrieben, das mit jeder Anlage mitgeliefert wird. Ausser einer eingehenden Beschreibung des Instrumentes wird auch eine Anzahl von Versuchen beschrieben, von denen einer das Messen der toten Zone und ihre Änderung mit Schwankungen der Verstärkung betrifft.

EE 77 753 *für weitere Einzelheiten*

Mikroanzeiger

Vertrieb: Kynmore Engineering Co. Ltd, 19 Buckingham Gate, London, W.C.2

(*Abbildung Seite 53*)

Der Mikroanzeiger MA-11 wird von Teledix, einer deutschen Tochtergesellschaft von Telefunken und Bendix, hergestellt und kann nunmehr von Kynmore Engineering Co Ltd bezogen werden.

Jeder Mikroanzeiger stellt die Ziffern 0...9 sowie einen Farbbalkensatz dar, jedoch sind auf Wunsch Darstellungen von Buchstaben, Symbolen und Vorzeichen lieferbar. Die weissen Zeichen auf schwarzem Hintergrund sind 3 mm breit und 5,6 mm hoch; sie werden auf eine mattierte Frontscheibe in schwarzem Kunststoffgehäuse projiziert, und zwar durch elf einzeln ausgerichtete Lampen in jedem Anzeiger. Die Aussenabmessungen der Anzeige-Einheit sind 8,6 mm breit, 13 mm hoch und 46,2 mm tief. Jede beliebige Anzeigerzahl kann nebeneinander für Dekaden und andere Anzeigen in Kassetten eingebaut werden. Schwarz eloxierte Standard-Leichtmetallkassetten für bis zu sechs Anzeiger sind lieferbar.

Der Mikroanzeiger selbst, d.h. der Baustein ohne Kassette, besteht aus drei Teilen (einer Fassung, einer Lampengruppe und der Anzeige), die fest zusammensitzen. Bei der Montage wird die Fassung normalerweise an die Rückseite der Kassette angeschraubt, die Lampengruppe und Anzeige zusammengeschraubt. Falls eine der Lampen ausfällt, können diese beiden Einheiten ohne Werkzeuge sofort nach vorn aus der Kassette herausgezogen werden. Anzeige und Ersatz-Lampengruppe werden dann wieder vorn in die Kassette eingeführt und mittels des Miniatursteckers in Drucktechnik in die Fassung eingesteckt. Die verschiedenen Teile werden in geeigneter Weise durch Federdruck fest und zwangsläufig zusammengehalten.

Die Wahl der logischen Schaltungen wird dem Benutzer überlassen; die meisten Typen transistorisierter Dekadenlogik eigen sich. Bei Betrieb mit 1,5 V und 70 mA (±5 mA für Langlebensdauer) ist die Leistungsaufnahme nur 0,1 W. Die Zeichen sind bei normalem Tageslicht bei Beobachtungswinkeln bis zu 30° sichtbar. Die Nennlebensdauer der Lampen ist 500 Stunden.

EE 77 754 *für weitere Einzelheiten*

Frequenzzähler

Hewlett-Packard Ltd, Dallas Road, Bedford

(*Abbildung Seite 53*)

Hewlett-Packard Ltd hat einen neuen, kompakten 5stelligen 2-MHz-Frequenzzähler 5534A eingeführt.

Messeinrichtungen sind vorhanden für Einzel- und Mehrfach-Periodendurchschnittsmessungen in Dekadenstufen bis zu 10^5 Perioden, impulsgesteuerte Zeitdauermessungen entweder mit der 100-kHz-Zeitbasisfrequenz des Gerätes oder extern angelegter Frequenz bis zu 2 MHz, Frequenzverhältnismessungen oder Mehrfache des Verhältnisses, sowie Zählen regelloser Ereignisse.

Kaltkathoden-Ziffernzählröhren geben eine klare und eindeutige Darstellung der Zähleranzeige. Anzeigespeicherung gibt eine kontinuierliche Darstellung des jüngsten Messresultats, selbst wenn das Gerät die nächste Messung vornimmt. Die Abfragezeit wird daher auf ein Minimum reduziert.

Hohe Empfindlichkeit und eine Eingangsimpedanz von 1 MΩ, 15 pF gestatten Nachweis und Messen von Signalen bis zu 100 V_{eff} herunter.

Die Bedienelemente sind einfach und zweckentsprechend; die Kommastellung ist automatisch, und das Gerät ist durchweg mit Halbleitern bestückt. Das Gehäuse ist für Tischarbeiten oder Gestelleinbau geeignet, und die Wartung des Instrumentes wird durch Steckkartentechnik erleichtert.

EE 77 755 *für weitere Einzelheiten*

Verzögerungs-Endlosbandgerät

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(*Abbildung Seite 53*)

Diese Ausrüstung wurde von Gresham Electronics Ltd für Aufnahme und Wiedergabe von Tonfrequenzsignalen oder Einschwingvorgängen auf einem endlosen Magnetband entwickelt.

Endlosbänder können in drei Längen von 381 cm, 285 cm und 190 cm benutzt werden; das Standard-Bandgerät hat zwei Geschwindigkeiten: 19 cm/s und 9,5 cm/s.

Bei der höheren Geschwindigkeit geben die drei Endlosbänder Umlaufzeiten von 20 Sekunden, 15 Sekunden und 10 Sekunden; die entsprechenden Zeiten für die niedrigere Geschwindigkeit sind 40 Sekunden, 30 Sekunden und 20 Sekunden.

Andere Bandgeschwindigkeiten und Verzögerungszeiten lassen sich durch einfaches Auswechseln des Antriebmotors sowie der Anzahl der Zwischenrollen erzielen.

Für 1" (25,4 mm)-Magnetband gibt es 14 Kanäle und für jeden derselben einen transistorisierten Aufnahmeverstärker- und Wiedergabeverstärkereinschub sowie ein umschaltbares Messgerät zur Überwachung der Aufnahme- und Wiedergabepegel.

Die Lösch- und Aufzeichnungsschaltungen kann man mit einem Schalter abschalten und damit ein komplettes Endlosband mit registrierten Signalen auf dem Magnetband "festhalten."

Die eingebaute Azimutgenauigkeit der Gresham-Aufnahme-, Wiedergabe- und Löschköpfe erlaubt, sie ohne mechanisches Nachjustieren auszutauschen.

EE 77 756 *für weitere Einzelheiten*

Wechsel-Gleichstrom-Vergleicher

Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1

(*Abbildung Seite 54*)

Es handelt sich hier um ein völlig neues, in sich geschlossenes Gerät mit Netzanschluss zum Messen der echten Effektivwerte von Wechselströmen und -spannungen mit einer Unsicherheit von ±0,05 Prozent über einen Frequenzbereich von 25 Hz...20 kHz.

Als Übertragungselement wird ein Einzel-Vakuum-Thermoumformer benutzt, der einen Wechselstrom-Gleichstrom-Übertragungsfehler von unter 0,01 Prozent hat; durch Schaltungskompensation können die Thermoumformer ohne Nacheichung ersetzt werden.

Ein grosser Vorteil des Gerätes ist die schnelle Kompensation der unbekannten Spannung oder des Stromes mittels einer regelbaren Gleichstromspannung, die dann ohne Übereilung gegen einen Bezugsgleichstrom gemessen werden kann, dessen Eichung die echten Effektivwerte des angelegten Signals gibt. Der Bezugsstrom wird einer zenerstabilisierten Stromversorgung entnommen, deren Ausgang sich bei Netzschwankungen von 10 Prozent weniger als 0,002 Prozent ändert.

Als externer Zubehör sind nur ein Normalelement und ein geeignetes Lichtzeiger-Galvanometer erforderlich, da das Gerät sonst mit eingebautem Spannungsteiler in sich geschlossen ist. Ein zweckdienliches Merkmal ist ein eingebautes Messgerät zur Bestimmung des ungefähren Wertes des unbekannten Signals vor Einstellung der Bedienelemente, was die Gefahr einer Überlastung des Vakuum-Thermoumformers verringert.

Das Instrument hat acht Spannungs- und Strombereiche, die die Werte von 0,5 bis zu 300 V bzw. von 5 mA bis zu 3 A überstreichen.

EE 77 757 *für weitere Einzelheiten*

Universal-Tester

Croydon Precision Instrument Co., Hampton Road, Croydon, Surrey

(*Abbildung Seite 54*)

Es handelt sich hier um ein kompaktes, vielseitiges Mehrzweck-Testgerät zum Messen von Widerständen, Gleich- und Wechselströmen sowie -spannungen.

Das Gerät besteht aus einem Rayleigh-Potentiometer mit vier Skalen, einem Thermoumformer, Spannungsteiler, Nebenschlusswiderständen, 50-Ω-Lichtmarken-Galvanometer und allen erforderlichen Schaltern.

Es ist tragbar und hauptsächlich für Werkstatt und Aussendienst geeignet, da es das unbequeme Zusammenschalten der für diese Messungen normalerweise erforderlichen Bauelemente überflüssig macht.

Die Geräteschaltungen können entweder aus zwei 2-V-Akkumulatoren oder zwei 2-V-Zenerdioden-Netzgeräten P3/S gespeist werden.

Es ist einfach zu bedienen und hat drei Bereichschalter für Spannung, Strom und Widerstand. Diese Schalter sind mechanisch verriegelt, um sicherzustellen, dass die nichtbenutzten Bereiche abgeschaltet sind.

Ströme von über 5 A können durch Anschalten externer Nebenschlüsse gemessen werden.

Im Eingangskreis liegt ein Suchmessgerät, das beim Messen von Gleich- oder Wechselspannungen bzw. -strömen die Bereicheinstellung erleichtert. Gleichspannungen bis zu 1,0999 V werden direkt mit dem Potentiometer vorgenommen, für höhere Spannungen wird bei Bereicheinstellung der Spannungsteiler automatisch eingeschaltet. Gleichstrommessungen werden mittels Potentiometer und Nebenschlüssen ausgeführt und die Bereiche durch Verstellen des Strombereichschalters umgestellt.

Der für Wechselstrommessungen eingesetzte Thermoumformer gibt den von der Wellenform unabhängigen wahren Effektivwert ab. Als Umformerelement wurde wegen seines niedrigen Umkehrfehlers ein Vakuum-Thermoumformer gewählt. Die Schaltungen sind so ausgelegt, dass die Messgenauigkeit von den Kenndaten des Vakuum-Thermoumformers völlig unabhängig ist.

Für Widerstandsmessungen wird das Rayleigh-Potentiometer als veränderlicher Zweig einer Wheatstoneschen Brücke benutzt. Die Brückenverhältnisse werden durch Schalter eingestellt.

Das Instrument wird in einen Teakholzkasten mit Metalldeckplatte eingebaut geliefert. Als Schalter findet die CROPICO-Type S.P.1 mit niedrigem Kontaktwiderstand, zwangsläufigem Rasten und hohem Isolierwiderstand Verwendung. Mit jedem Gerät kommen ausführliche Betriebsanweisungen.

Insgesamt sind 40 Bereiche vorgesehen, und zwar von 0 ... 1 MΩ, von 0 ... 600 V– und 2 ... 600 V~ (25 Hz ... 10 kHz), von 0 ... 5 A– und 5 mA ... 5 A~.

EE 77 758 *für weitere Einzelheiten*

Thyristortreiber-Baustein

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

(*Abbildung Seite 54*)

Der von Mullard Ltd angekündigte Thyristortreiber-Baustein MY5000 wurde zur Vereinfachung der Thyristor-Verwendung auf vielen allgemein üblichen elektrischen Anwendungsgebieten entwickelt. Er ist der einfachste Baustein einer Serie und kann gleichzeitig bis zu zwei 70-A-Thyristors treiben, z.B. als zwei Zweige einer einphasigen Brückenschaltung.

Für hochgezüchtete, sehr genaue Motordrehzahlenregelung mit komplexer Rückführung sind zwar kompliziertere Treiberbausteine erforderlich, jedoch ermöglicht dieser einfache Baustein eine zufriedenstellende und wirtschaftliche Lösung für viele einfachere Anwendungsmöglichkeiten in der Drehzahlregelung, Ofenregelung und Lichtregelung, sowie auch für die Steuerung von Gegentakt-Thyristoranordnungen.

Thyristor-Benutzer finden, dass die Erzeugung eines zuverlässigen Treiberimpulses, der den Thyristor fehlerfrei zündet und auch die Lage des Impulses mittels eines Regelelementes oder einer Rückführung steuert, Schwierigkeiten bereitet. Jede Anwendungsmöglichkeit erfordert selbstverständlich ihre eigene Rückführungsanordnung; bei Einsatz des Mullard-Bausteins, der Thyristor-Steuerimpulse innerhalb zweckmässiger Toleranzen abgibt, sind nur die Gleichstrom- und Rückführungsprobleme zu lösen. Die häufiger auftretenden dieser Probleme werden in den Mullard Technical Reports behandelt werden.

Der Trigger-Baustein MY5000 ist gegen Umgebungseinflüsse durch komplette Kunststoffeinkapselung geschützt, die auch Schaltungselemente und Verbindungen gegen Beschädigung durch mechanische Schwingungen oder Stoss schützt.

EE 77 759 *für weitere Einzelheiten*

Laufzeitspeicher

The M.E.L. Equipment Co. Ltd. 207 Kings Cross Road, London, W.C.1

(*Abbildung Seite 54*)

Aus vereinheitlichten Teilen zusammengebaute neue ultrasonische Verzögerungsleitungen können wirtschaftlich für die Serienspeicherung von Information bis zu 1000 Bits eingesetzt werden.

Jede dieser als "Verzögerungs-Bausteine YL2108" aufgeführten Einheiten wird komplett mit magnetostriktivem Drahtverzögerungselement sowie Festkörper-Eingangs- und Festkörper-Ausgangsschaltungen geliefert. Jede beliebige Verzögerung von 100 μs bis zu 3,2 ms ist mit einem Regelbereich von 20 μs lieferbar.

Die Bausteine können im Temperaturbereich von −10°C ... +60°C mit Bit-Folgen bis zu 350 kHz betrieben werden. Eine 12-V-Stromversorgung, die positiv, negativ oder symmetrisch sein kann, ist erforderlich. Die Bausteine sind daher für die meisten mit p-n-p- oder n-p-n-Transistoren bestückten Logiksysteme geeignet und mit den MEL-Digitalschaltungsblöcken der Serie 1 völlig kompatibel.

Die Verzögerungsleitungen arbeiten üblicherweise nach der zyklischen Arbeitsweise, nach der der Ausgang verstärkt und in den Eingang zurückgespeist wird, so dass die Information gespeichert bleibt, solange die Stromversorgung eingeschaltet ist. Die Information kann jederzeit entnommen werden, wobei die Höchstzugriffzeit der Gesamtverzögerung in der Leitung gleich ist.

In den einfacheren Datenverarbeitungsgeräten sollten für diese Bausteine viele Anwendungsmöglichkeiten bestehen. Die Leitungen können als kurzzeitige Speicher in kleineren Rechenmaschinen, Fakturiermaschinen, prozessgekoppelten Elektronenrechnern und ähnlichen Anwendungsmöglichkeiten dienen. Sie können auch zusammen mit Dauerspeichern wie z.B. Kernspeichern Verwendung finden. In diesem Fall speichern die Leitungen Information für kurze Zeit, während Berechnungen oder Sortiervorgänge ablaufen.

Die Leitungen sind in Gehäusen aus plattiertem Stahl mit Abmessungen von 235 × 171 × 38 mm untergebracht. Sie können in jeder Lage oder in dem MEL-Einsatz- und Gestellsystem eingebaut werden. Eine der schmalen Seiten dient als Frontplatte, auf der die Ausgangstestpunkte und Verzögerungsregelung angeordnet sind. Anschlüsse für Stromversorgung, Signaleingang und -ausgang sind auf der Rückseite angeordnet.

EE 77 760 *für weitere Einzelheiten*

Rauschklirrmessplatz

Marconi Instruments Ltd, St. Albans, Hertfordshire

(*Abbildung Seite 55*)

Marconi Instruments Ltd hat einen neuen volltransistorisierten Rauschklirrmessplatz OA2090 zum Testen von Kabel- und Mehrkanal- Richtfunkverbindungen mit bis zu 2700 Kanälen Kapazität entwickelt, der den C.C.I.R.-Empfehlungen entspricht. Er besteht aus zwei Geräten und ist viel kleiner und leichter als seine Vorgänger.

In Mehrkanalsystemen ist es unbedingt

erforderlich, die in jedem Kanal auf Verkehr in anderen zurückzuführende Kreuzmodulationsstörung so gering wie möglich zu halten. Dieses Nebensprechen wird hauptsächlich durch nichtlineare und Phasenverzerrungen hervorgerufen und wird als weissem Rauschen ähnliche Störung hörbar, die man durch Messen eines Rauschleistungsverhältnisses bestimmen kann.

Der Messplatz arbeitet nach den folgenden Prinzipien: Weisses Rauschen wird mit zweckmässiger Bandbreite und Leistungspegel dem System zugeführt, das sehr genau ein vollbelastetes Mehrkanal-Fernsprechsystem simuliert. Durch Einschalten eines Filters mit engem Sperrbereich zwischen Rauschquelle und System wird ein geräuschfreier Kanal erzeugt. Ein auf den geräuschfreien Kanal abgestimmter Empfänger wird benutzt, um den durch Kreuzmodulation der Komponenten des weissen Rauschens im übrigen Frequenzband hervorgerufenen Rauschpegel mit dem Originalsignal des Kanals zu vergleichen.

Der Rauschgenerator ist mit einer Halbleiter-Spezialdiode als Rauschquelle bestückt. Er nimmt bis zu neun Schnellwechselfilter auf, und zwar bandbegrenzende wie auch Sperrband-Typen. Die Filter enthalten farbgekennzeichnete Kippschalter und sind alle vor dem Leistungsmonitor angeordnet; die Rauschbandgrenzen und die Lage des gesperrten Bereiches können daher durch einfaches Umlegen von Schaltern bestimmt werden, während der Leistungspegel laufend überwacht wird. Die Belastung des Systems ist mittels eines im Ausgang liegenden Abschwächers regelbar.

Der Empfänger, in dem sich bis zu sechs Festfrequenzen vorsehen lassen, ist auch mit auswechselbaren Filtern bestückt, die den Sperrbandfilterfrequenzen des Generators entsprechen. Ein Eingangsabschwächer ist so geeicht, dass man das Rauschleistungsverhältnis in Dezibel direkt ablesen kann.

Der Messplatz, der auch für Gestelleinbau lieferbar ist, hat folgende Aussenmasse: etwa 190 × 470 × 433 mm; er wiegt ungefähr 12 kg.

EE 77 761 *für weitere Einzelheiten*

Ferrit-Topfkerne

Standard Telephones & Cables Ltd. Magnetic Materials Division, Edinburgh Way, Harlow, Essex

(*Abbildung Seite 55*)

Die Magnetic Materials Division der STC hat ihr Fertigungsprogramm durch Aufnahme einer Anzahl neuer Topfkerne nach internationalen Normen erweitert. Ein Konstruktionsmerkmal dieses Kerntyps ist die Abstimmvorrichtung, die gänzlich innerhalb des Kernes liegt. Das ermöglicht in Geräten die kleinstmögliche Raumbeanspruchung für den Kernaufbau und gewährleistet Stabilität unter den schwierigsten Umgebungsbedingungen.

Die Kerne sind in drei Grössen und zwei Werkstoffen lieferbar, und zwar Grössen 26, 18 und 14 mm und Werkstoffen SA501 und SA503. Von Zeit zu Zeit wird die Fertigung weiterer Grössen in verschiedenen Werkstoffen aufgenommen werden, um dann ein geschlossenes Programm für alle Anwendungsmöglichkeiten anbieten zu können.

Die Werkstoffe SA501 und SA503 sind verlustarme, hochstabile Arten, die für Schwingkreise bis zu 500 kHz und für nichtabgestimmte Kreise bis zu ungefähr 1 MHz geeignet sind. SA501 hat einen sehr niedrigen Temperaturfaktor und ist für Verwendung mit Glimmerkondensatoren bestimmt. SA503 hat einen kontrollierten positiven Temperaturbeiwert und ist für Verwendung mit Polystyrolkondensatoren gedacht. IEC-Kerne sind sowohl für Chassismontage, als auch für gedruckte Schaltungen sehr geeignet. Für alle Grössen sind Delrin-Spulen von STC erhältlich.

Während der Herstellung der IEC-Kerne wird bei STC die chemische Zusammensetzung und Reaktivität laufend sorgfältig überwacht. Besondere Aufmerksamkeit wird der Kontrolle der Mikrostruktur des Ferrits gewidmet, und das Verfahren soll Kerne ergeben, die nicht ihresgleichen haben.

EE 77 762 *für weitere Einzelheiten*

Betriebsstunden-Anzeiger

Venner Ltd, Kingston By-Pass, New Malden Surrey

(*Abbildung Seite 55*)

Der von Venner Ltd herausgebrachte, durch einen Synchronmotor getriebene Betriebsstunden-Anzeiger SHR1 kann mit jedem Gerät eingesetzt werden, das mit Wechselstrom geregelter Frequenz betrieben wird. Das Instrument zeigt Stunden in Zehnteln und bis zu 99 999,9 Stunden an und wird zum Weiterzählen automatsich auf Null zurückgestellt.

Anwendungsmöglichkeiten bestehen in Lebensdauerprüfung, Materialverbrauch, Laufzeit von Pumpen und elektronischen Geräten, sowie für die Anzeige des Zeitpunktes, zu dem Wartung und Überholen von Drehautomaten und ähnlichen Werksausrüstungen fällig wird.

Es gibt zwei Typen, und zwar für Aufputzmontage mit Kabel- und Rohreinführung und für bündigen Einbau. Ein vierpoliger Anschlussblock gestattet Betrieb mit drei verschiedenen Wechselspannungen: 100 ... 120 V, 200 ... 250 V und 380 ... 460 V, 50 Hz. Die Leistungsaufnahme bei 240 V ist 2,2 W. Abmessungen der Einbauausführung sind ungefähr 76 mm breit, 70 mm tief und 51 mm hoch.

Auf Wunsch kann das Instrument auch für das Registrieren von Minuten anstelle von Stunden geliefert werden.

EE 77 763 *für weitere Einzelheiten*

Temperatursonde

Vertrieb: The Solartron Electronic Group Ltd. Victoria Road, Farnborough, Hampshire

(*Abbildung Seite 56*)

Die Temperatursonde ist ein tragbares thermoelektrisches Instrument, das in kleinen Bauelementen Temperaturänderungen hervorrufen kann. In eine mit auswechselbaren Spitzen ausgerüstete Sonde sind zwei thermoelektrische Kühlelemente eingebaut. Auf der Frontplatte eines robusten Metallgehäuses, das eine Gleichstromversorgung und das Wärmeaustauschsystem enthält, sind einfache Bedienelemente und ein grosser Temperaturanzeiger angeordnet.

Das Gerät erhöht oder senkt die Temperatur einzelner eingebauter Bauelemente und hilft auf diese Weise bei der Prototypkonstruktion und -entwicklung. Sondenspitzen sind in verschiedenen, für maximalen thermischen Kontakt mit den zu untersuchenden Bauelementen ausgelegten Formen lieferbar. Weitere Anwendungsmöglichkeiten bestehen in der Abnahmeprüfung für Zulieferungen, Fehlersuche und Schrumpfmontage kleiner Teile durch direktes Erwärmen oder Kühlen. Auch in den Forschungslaboratorien der Chemie und Medizin werden sich viele Anwendungsmöglichkeiten ergeben.

Der Temperaturbereich der Temperatursonde ist —20° C...+70° C, und 60 Prozent der Endtemperatur werden in zwei Minuten Ansprechzeit erreicht.

In der "Temperatursonde TP-10" pumpen zwei thermoelektrische Kühlelemente Wärme zwischen der Testfläche und der Umgebungstemperatur. Beim Kühlen wird dem Prüfling (z.B. einem elektronischen Bauelement in einer aktiven Schaltung) durch die speziell geformten Spitzen Wärme entzogen und durch Leitung auf die Oberflächen der thermoelektrischen Elemente übertragen. Durch Anlegen eines Gleichstroms wird die Wärme auf die gegenüberliegenden Flächen der Elemente gepumpt und durch Kühlflüssigkeit abgeleitet. Die Flüssigkeit wird durch einen Flüssigkeit-Luft-Wärmeaustausch zirkuliert und die Wärme an die Atmosphäre abgegeben. Beim Erwärmen wird der Wärmefluss durch Umpolen des Gleichstroms umgekehrt. Der Atmosphäre entzogene Wärme wird in die Testzone gepumpt. Die gepumpte Wärmemenge ist eine Funktion des Gleichstrompegels. Die Temperatur wird mittels eines Kupfer - Konstantan - Thermoelementes abgetastet und an einem Weston-Pyromillivoltmeter gemessen.

Das Instrument wird von Daystrom Ltd in Kanada gebaut und kann in Grossbritannien von The Solartron Electronic Group Ltd bezogen werden.

EE 77 764 *für weitere Einzelheiten*

Magnetischer Miniaturmesswertgeber

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucester

(*Abbildung Seite 56*)

Advance Controls Ltd hat das bestehende Fertigungsprogramm durch einen neuen magnetischen Miniaturmesswertgeber MT3SR erweitert. Diese Geber enthalten keine beweglichen Teile und erzeugen Impulse von umlaufenden Wellen und Zahnrädern, ohne dieselben zu berühren. Diese Impulse können Drehzahlwächter, Geschwindigkeitsschalter, analoge oder digitale Tachometer, Ratiometer sowie Anzeiger für Geschwindigkeitsdifferenzen treiben und haben noch weitere zahlreiche Anwendungsmöglichkeiten in der Industrie.

Der Messwertgeber wird so angebaut, dass seine Polschuhe nahe der rotierenden Verzahnung stehen; als Abstand wird 0,76...1,5 mm empfohlen, wobei etwas Lagerspiel berücksichtigt ist. Selbst bei Umfangsgeschwindigkeiten von nur 50 mm/s werden dann Impulse mit über 300 mV Amplitude erzeugt.

Der neue Messwertgeber ist hermetisch dicht in ein robustes, für fast alle industriellen Bedingungen geeignetes Messinggehäuse eingebaut, das 51 mm lang ist und zur Erleichterung der Montage ein 5/8"-Schrauben-Feingewinde (BSF) hat. Zwölf andere Modelle für verschiedene Verwendungszwecke sind sofort greifbar, und mehr als 60 Sonderausführungen können geliefert werden.

EE 77 765 *für weitere Einzelheiten*

Hohlleiterschalter

The Marconi Co. Ltd, Chelmsford, Essex

(Abbildung Seite 56)

Marconi Ltd kann nunmehr einen völlig elektronischen Hohlleiterschalter liefern, der mit 2 W belastbar ist und den Frequenzbereich 7,75 ... 8,5 GHz überstreicht. Der für Mikrowellen-Nachrichtenverkehr entwickelte Schalter hat eine Dämpfung von 50 dB bei wesentlich verbesserter Zuverlässigkeit, Schaltgeschwindigkeit, Raumeinsparung und Leistungsfähigkeit im Vergleich mit älteren mechanischen Schaltmethoden.

Der neue Schalter ist für Hohlleiter Nr.15 ausgelegt und besteht aus einer Hohlleiterlänge in Sonderausführung mit Ferriteinsatz und einer um die Aussenseite des Hohlleiters gewickelten Spule, die ein magnetisches Längsfeld erzeugt. Wenn dieses Feld erregt wird, steigt die Dämpfung schnell auf 50 dB. Mit einer transistorisierten Schalteinrichtung lässt sich eine Schaltzeit von unter 1 ms erzielen.

EE 77 766 *für weitere Einzelheiten*

Tunneldioden-Stromversorgung

Coutant Electronics Ltd, 3 Trafford Road, Richfield Estate, Reading, Berkshire

(Abbildung Seite 56)

Die verhältnismässig niedrigen Spannungen für Tunneldiodenschaltungen erfordern eine Spezialtechnik, vor allem wenn es sich um Stromstärken von mehreren Ampere handelt.

Coutant Electronics Ltd hat eine Konstantstromquelle entwickelt, die gedruckten Standardbandleitungen angepasst ist. Die Ausgangsverbindungen werden durch direktes Anlöten der Bandleitungen an die zweiseitigen Druckschaltungen der Stromversorgungen hergestellt, wobei für Kontinuität in der Masseebene Vorsorge getroffen ist.

Die Ausgangsspannung ist bei 8 A Stromstärke kontinuierlich zwischen 0 und 100 mV regelbar, und für die zugehörigen Transistorschaltungen sind zwei Hilfsversorgungen für +5 V und -5 V bei 6 A vorhanden.

Fehlersignale werden durch Miniatur-Koaxialkabel und -Verbindungen von der Last zum Steuerverstärker zurückübertragen. Die Ansprechzeit des Konstanthalters auf steile Änderungen des Laststromes ist näherungsweise das erreichbare Minimum und hängt hauptsächlich von der Länge der Bandleitung zwischen Stromversorgung und Messpunkt ab.

Die Stromversorgung ist durchgehend mit Silizium-Halbleitern bestückt und kann in Umgebungstemperaturen bis zu 60° C betrieben werden. Sie ist für Standard-19"-Gestelle konstruiert und hat eine Feldhöhe von etwa 133 mm.

EE 77 767 *für weitere Einzelheiten*

Digitalfrequenzmesser

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 56)

Mit der Einführung des Digitalfrequenzmessers 719A kann Dawe Instruments Ltd nunmehr die Nachfrage nach Instrumenten, die bei mässigem Aufwand umfangreiche Allgemeinverwendung in Labor und Industrie finden, decken. Mit fünfstelliger Leuchtanzeige und quarzgesteuertem 10-kHz-Oszillator hat der Frequenzmesser viele Einsatzmöglichkeiten, z.B. für das Messen von Frequenzen, Perioden und Drehzahlen, sowie das Zählen regelloser Impulse.

Der 719A beruht auf dem üblichen Digitalzähler-Zeitgeber-Prinzip: die elektrische Eingangswellenform wird verstärkt und begrenzt, so dass eine Impulsreihe entsteht, die während einer Durchlassperiode gezählt werden kann oder selbst zur Steuerung der Torschaltung benutzt wird. Für Frequenzmessungen schliessen oder öffnen die Taktimpulse des Quarzoszillators in Verbindung mit dekadischen Teilern die Torschaltung für eine vorgewählte Periode zwischen 0,01 und 10 s, während der die Eingangsimpulse gezählt und angezeigt werden. Drehzahlen werden auf dieselbe Weise mit einer Durchlassperiode von 6 s gemessen und direkt in UPM angezeigt. Für Periodenmessungen ist das Verfahren umgekehrt, und die Anzahl der vom Quarzoszillator abgegebenen Impulse (oder eine Dekadenteilung dieser Anzahl) wird während einer Periode des Eingangssignals gemessen. Man kann das Gerät auch für "Stoppuhr"-Arbeiten einsetzen, und in diesem Fall wird die Anzahl der Impulse für eine festgelegte Eingangszeit gezählt. Regellose Impulse können auch gezählt werden, vorausgesetzt dass der Abstand zwischen aufeinanderfolgenden Impulsen nicht unter 0,1 μs liegt.

Frequenzen können zwischen 0 und 10 MHz, Perioden und regellose Impulse bis zu 99999 s gemessen werden. Die Gesamtgenauigkeit ist ± 1 Zählimpuls $\pm$ der Quarzgenauigkeit, die bei Zimmertemperatur auf 1×10^{-6} eingestellt wird. Die Ziffernanzeige hat Kommaautomatik und eine Anzeige "Zählung überschritten"; die Endanzeige bleibt für eine vorwählbare Zeitspanne zwischen 1 und 15 s sichtbar, wenn sie nicht manuell gelöscht wird. Transistorisierung und ein neuer dekadischer Teiler, für den ein Patent angemeldet wurde, haben die Schaltung wesentlich vereinfacht.

Das Modell 719A ist für vielseitigen Einsatz geeignet, z.B. in der Tachometrie, Frequenzeichung, Relaiszeitprüfung, synchronisierten Stoppuhr-Aufgaben, Phasen- und Drehmomentmessungen, sowie vor allem dort, wo bisher ein Digitalfrequenzmesser als zu aufwendig betrachtet wurde.

Die Leistungsaufnahme ist nur 15 W (Netz) und das Gewicht etwa 16 kg. Die Abmessungen des für Tischaufbau oder Gestelleinbau geeigneten Gehäuses sind ungefähr 520 × 280 × 305 mm tief.

EE 77 768 *für weitere Einzelheiten*

Funktionsverstärker

Bendix Electronics Ltd, High Church Street, New Basford, Nottingham

(Abbildung Seite 57)

Bendix Electronics Ltd kündigt einen neuen chopperstabilisierten Mehrzweck-Festkörperfunktionsverstärker 470 an. Mit Hilfe von Gegenkopplung im Verstärker wurde eine sehr präzise Verstärkungskennlinie gewonnen, die nur eine Funktion der Gegenkopplungsimpedanz ist. Ausserdem hat der Eingang praktisch Zeropegel, so dass kein Strom in den Verstärker fliesst, wodurch dieser sich in idealer Weise für das Summieren von Eingangsströmen eignet. Zur Erzielung höchster Zuverlässigkeit ist das Modell 470 mit Silizium-Epitaxial-Planar-Transistoren bestückt. Es ist für einen Temperaturbereich von 0...70° C ausgelegt und hat bei Gleichstrom eine Verstärkung von 8×10^6, die bei 2 MHz auf 1 abfällt.

Als Hilfsverstärker hat das Modell eine Verstärkung von 2000, die von 0,045 Hz an um 20 dB/Dekade abfällt, und als Breitbandverstärker eine solche von 4000, die über 1 kHz um 20 dB/Dekade abfällt. Die Eingangsimpedanz ist 6 kΩ und die Ausgangsspannung 10 V an 1 kΩ Abschluss.

EE 77 769 *für weitere Einzelheiten*

Prozess-Zeitgeber

Vertrieb: Rayleigh Instruments Ltd, 271 Kiln Road, Thundersley, Essex

(*Abbildung Seite 57*)

Ein neues Sortiment der von Dold & Söhne in der Bundesrepublik Deutschland hergestellten hochpräzisen Zeitgeber und Verzögerungsrelais ist nunmehr von Rayleigh Instruments Ltd erhältlich.

Diese von Synchronmotoren getriebenen Instrumente erprobter Bauart genügen den sehr strikten Pflichtenblättern für Zeitgenauigkeit und elektrische Kenndaten. Sie werden in verschiedenen Ausführungen hergestellt, von denen einige bis zu fünf verschiedene Zeitbereiche haben können. In den Mehrbereich-Modellen erfolgt die Bereichwahl durch Verstellen einer Steuerschraube. Verzögerungszeiten lassen sich für jede Zeit zwischen den Grenzen O,1 s und 120 h einstellen, die in den Bereich des gewählten Types fällt.

Die Instrumente können entsprechend Kundenwünschen mit entweder bei Beginn oder Ende des Zyklus betätigten Kontakten bestückt werden.

EE 77 770 *für weitere Einzelheiten*

Automatischer Leistungsregler

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(*Abbildung Seite 57*)

Kent Precision Electronics Ltd hat die Fertigung einer Serie von automatischen Festkörper-Reglern modernster Konstruktion für Verwendung mit Stellmotor-Reglern aufgenommen. In der normalen automatischen Regelung wird Steigern und Senken der Ausgangsleistung durch Schalten des Stellmotor-Reglers erzielt.

Geräte der Serie KPE arbeiten mit einem geeichten Festkörper-Standardsteuerinstrument für proportionale Steuerung eines Gleichstrom-Kollektormotors. Für grosse Fehler wird der Motor voll eingeschaltet und innerhalb des Proportionalbereiches von 2 Prozent verlangsamt, bis er ohne Überschiessen oder Pendeln beim Sollwert zur Ruhe kommt. Für alle normalen Anwendungszwecke kann man ohne Schwierigkeiten eine tote Zone von 0,5 Prozent erzielen. Wo für allgemeine kommerzielle Stellmotor-Regler eine herkömmliche "Ein-Aussteuerung" erforderlich ist, kann ein Standard-Gerät KPE geliefert werden, dessen geeichtes Festkörper-Steuerinstrument über ein Standardrelais ein- und ausschaltet.

Durch Zuführung sekundärer Funktionseingänge kann das geeichte Festkörper-Standardsteuerinstrument programmiert arbeiten, wenn konstantgehaltene Stromdichten erforderlich sind—z.B. in Bädern für galvanische Metallabscheidung. Ein Standard-Stromdichte-Konstanthalter ist in der Abbildung illustriert.

Das Gerät ist als robustes Schalttafelinstrument oder für Gestelleinbau lieferbar. Da keine beweglichen Teile vorhanden sind, zeichnen sich automatische Regelsysteme mit KPE-Instrumenten durch zuverlässigen und wartungsfreien Betrieb aus.

EE 77 771 *für weitere Einzelheiten*

Automatischer Abschwächer

Sangamo Weston Ltd, Great Cambridge Road, Enfield, Middlesex

Bei Sangamo International ist nunmehr die Fertigung des automatischen Abschwächers SA-2L angelaufen. Dieses Gerät ist hauptsächlich für Anwendungsgebiete wichtig, in denen bei der Aufbereitung analoger Daten gelegentlich Spitzen oder unerwartete Amplituden auftreten, die Anzeige- oder Registriergeräte wie Magnetbandgeräte, Oszillografen, Streifen- oder Koordinatenschreiber übersteuern.

Modell SA-2L schwächt Signale, die einen einstellbaren Triggerpegel überschreiten, um einen bekannten Wert und erweitert dadurch den dynamischen Bereich jedes Systems mit konstantem Rauschen. Anwendungsmöglichkeiten bestehen jedoch nicht nur, wo es sich um Spitzen handelt. Der Abschwächer kann auch als Bereichdehner eingesetzt werden und gibt dann volle Systemgenauigkeit bei zwei einstellbaren Punkten.

EE 77 772 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein X-Band-Spektralanalysator und Wobbelgenerator von R. W. Eldred

Zusammenfassung des Beitrages auf Seite 2-6

Dieser Beitrag beschreibt kurz einen kombinierten Spektralanalysator und Wobbelgenerator für das X-Band. Der Wobbelgenerator ergibt sich aus der zur Gewährleistung konstanter Empfindlichkeit über den elektronischen Wobbelbereich im Spektralanalysator angewandten Methode. Ausserdem wird ein Verfahren zur Verwendung eines internen Absorptionswellenmessers in der AFN-Schleife zur Konstanthaltung der Hilfsoszillatorfrequenz für "manuellen Durchlauf" beschrieben.

Versuche mit magnetohydrodynamischer Energieerzeugung von G. J. Womack

Zusammenfassung des Beitrages auf Seite 7-14

Dieser Beitrag berichtet über Resultate, die mit einem kleinen, durch gasförmige Propan-Sauerstoffverbrennung getriebenen MHD-Generator erzielt wurden. Ein verhältnismässig grosser Dauermagnet, der eine konstante Kraftliniendichte von 9800 Gauss erzeugt, wurde benutzt. Die Verbrennungsprodukte wurden mit bis zu etwa 1,6 Gewichtsprozent Kalium geimpft. Versuche wurden sowohl mit heissen Kohleelektroden, als auch mit wassergekühlten Kupferelektroden durchgeführt.

Die elektrische Leitfähigkeit der Arbeitsflüssigkeit wurde mit beiden Elektrodentypen durch Messen der I-V-Kennlinien bei erregtem wie auch angelegtem elektrischen Feld gemessen. Zur Unterstützung der Auswertung dieser Ergebnisse wurde ein Zweistrahl-Natriumlinienumkehrapparat gebaut, um die Gastemperatur zu messen. Für heisse und kalte Elektroden wurde gute Übereinstimmung zwischen den gemessenen und theoretisch berechneten Werten der elektrischen Leitfähigkeit erzielt. Für die kalten Elektroden war die Leitfähigkeitsform elektronisch und nicht ionisch, so dass die kalte Elektrode Elektronen emittieren muss. Der exakte Mechanismus dieser Emission ist noch nicht voll verständlich, ist aber wahrscheinlich auf den Kohlenstoffniederschlag auf der Elektrodenoberfläche zurückzuführen. Dieser Kohlenstoffniederschlag ist anscheinend strukturlos und sehr porös bei niedriger thermischer Leitfähigkeit. Diese Schicht wird sich daher dem Staupunkt des Gasstroms nähern, was für hohe thermonische Emission ausreichend wäre.

Aus den Messungen der elektrischen Leitfähigkeit können die Kollisionsquerschnitte ermittelt und mit den nach Frost (14) berechneten theoretisch vorausgesagten Werten verglichen werden. Annehmbar gute Übereinstimmung wurde erzielt.

Niederfrequentes Spannungsrauschen wurde beobachtet.

Zur Bestimmung von $w_e\tau_e$ wurden Hallgeneratormessungen durchgeführt.

Ein Eingangsübertrager mit niedrigen Erdschlussströmen von D. G. Wyatt

Zusammenfassung des Beitrages auf Seite 16-19

Die durch Eingangsübertrager in elektrischen Umschaltungen eingebrachten Fehlerquellen werden untersucht und der Entwurf eines Übertragers mit niedrigen Erdschlussströmen eingehend beschrieben. Besondere Aufmerksamkeit wird der angewandten Neutralisierungsschaltung gewidmet. Die Kenndaten eines kompletten Übertragers werden gegeben.

Ein preiswerter Mittelwertrechner für die direktgekoppelte Analyse neurophysiologischer Daten von D. Burns, W. Ferch und G. Mandl

Zusammenfassung des Beitrages auf Seite 20-24

Steigendes Interesse am Zeitdurchschnittsverhalten von Nervenzellen hat zu einem wachsenden Bedarf an Instrumenten, die schnelle, direktgekoppelte Klassierung von Zeitintervallen gestatten, geführt. Ein verhältnismässig preiswerter Sonderzweckrechner wird beschrieben, der sowohl die Intervallanalysen wie auch Digital-, Kreuz- und Autokorrelation durchführen kann.

Multivibratoren ohne Phasenverschiebung von A. Pugh

Zusammenfassung des Beitrages auf Seite 24-25

Untersuchungen an Schaltungen, die positiv und negativ verlaufende Impulse verarbeiten, haben zur Konstruktion von Transistor-Multivibratorschaltungen mit Basisschaltungsstufen anstelle von Emitterschaltungsstufen geführt. Der Beitrag macht auf die ziemlich neuartigen Eigenschaften einer Reihe nach den obigen Prinzipien entworfener Schaltungen aufmerksam und weist auf einige Anwendungsmöglichkeiten hin.

Hochstabiles NF-Bandfilter von P. G. Simpson

Zusammenfassung des Beitrages auf Seite 26-31

Der Entwurf eines sehr hochstabilen Tiefstfrequenz-Bandfilters, das eine Phasenkonstanz von besser als 1×10^{-10} hat, wird beschrieben. Die theoretischen Gesichtspunkte für die Wahl der gekoppelten Systeme werden beschrieben und einige präzise Dämpfungs- und Phasenwinkelkurven zusammen mit einer kurzen Erläuterung ihrer Ableitung gegeben. Einige praktische Konstruktionsmerkmale kompletter Filter werden erwähnt.

Zwei ausgeprägte Grenzen für Transistor-Rückkopplungsgeneratoren von B. Shahzadi

Zusammenfassung des Beitrages auf Seite 32-34

In der Literatur kann man für ungedämpfte Schwingungen zwei verschiedene Kriterien finden. Autoren weisen im allgemeinen auf ein Kriterium hin und lassen das andere versehentlich ausser acht.

In diesem Beitrag werden die Kriterien für mehrere Typen der Rückkopplungsgeneratoren ausgearbeitet und diskutiert.

Ein kombinierter Spannungs- und Erdschlussanzeiger für Gleichstromversorgungen von G. T. Edwards

Zusammenfassung des Beitrages auf Seite 34-35

Eine einfache Schaltung für Verwendung mit Gleichstromversorgungen wird beschrieben. Sie zeigt die Ausgangsspannung an und ausserdem — falls eine der Klemmen an Masse liegt — welche Klemme geerdet ist. Die Einrichtung ist vor allem für Anlagen geeignet, in denen eine Stromversorgung mehrere entfernt gelegene Anschlüsse speist.

Vorregelung einer transistorisierten Stromversorgung mit steuerbaren Siliziumgleichrichtern von D. Ophir

Zusammenfassung des Beitrages auf Seite 36-40

Ein zuverlässiger Vorregler mit steuerbarem Siliziumgleichrichter für eine transistorisierte Stromversorgung wird beschrieben. Der Vorregler wird bei hoher Frequenz von einer Gleichstromquelle betrieben.

Umkehrbarer Dekadenzähler mit Rückstellung durch Gleichstromlogik von J. H. Smith

Zusammenfassung des Beitrages auf Seite 41-43

Ein zuverlässiger, umkehrbarer Dekadenzähler mit zwangsläufiger Gleichstromtor-Rückstellung wird beschrieben. Der Zähler ist so konstruiert, dass er in jeder Zählrichtung denselben Ausgangscode abgibt. Ausserdem wird die Entwicklung einer ähnlichen Dekade beschrieben, deren Dezimalausgabe für beide Zählrichtungen im 8-4-2-1-Binärcode erfolgt. Diese Zähler sind nicht selbstkomplementierend.

Arbeitsweise eines magnetisch gekoppelten Multivibrators mit Kernen mit rechteckiger Hystereseschleife von G. Smiljanic

Zusammenfassung des Beitrages auf Seite 44-46

Einige Eigenschaften eines frequenzgesteuerten, magnetisch gekoppelten Multivibrators werden unter der Voraussetzung besprochen, dass der Treiber-Übertrager Kerne mit rechteckiger Hystereseschleife hat. Es wird gezeigt, dass die Frequenz des Multivibrators linear von der Eingangsspannung abhängt. Die Frequenz kann durch variable Widerstände geändert, durch Zener-Dioden gegen Schwankungen der Eingangsspannung oder Temperatur stabilisiert und durch Regelung des Gleichstroms über einen breiten Bereich variiert werden, wobei die Amplitude der Rechteckwellenform des Ausgangsimpulses konstant bleibt.