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Commentary

THE Royal Society this year celebrates the three-hundredth anniversary of its foundation and to mark this great historic occasion a programme of Tercentenary Celebrations was held from 18 to 26 July. In a number of ways the Society is unique: it is certainly the oldest scientific society in the world which has enjoyed a continuous existence and unlike the Académie des Sciences of France or the Akademia Nauk of Russia it is not a government institution, though it has many ties with the government and in some cases acts for it.

Although justly proud of its long and fine traditions the Royal Society is by no means a slave to them and it is, indeed, interesting to recall that it was, so to speak, a rebellion against tradition which brought about its inception. In the early seventeenth century it was widely held that the writings of the great classical philosophers were the one source of wisdom from whence all knowledge flowed: so far as scientific matters were concerned this meant, largely, the writings of Aristotle. The works of such men as Galileo the founder of mechanics, Gilbert, the founder of experimental physics and of Kepler on the laws governing planetary motions, found no place in the teachings at the universities. It is to a small group of men led by John Wallis, the mathematician, John Wilkins and the medical doctors Glisson, Ent, Goddard and Merret, who recognized the importance of the works of Galileo and his contemporaries and, more important realized the immense value of the point of view which they represented, namely, that systematic observation and experiment were the proper means of investigating natural happenings, that the Royal Society owes its foundation. At first this group met for discussions in London but about 1648 Wilkins, Wallis and Goddard removed to Oxford where they continued their scientific discussions in Wilkins' rooms in Wadham College. In 1659 the Oxford and London groups, reinforced by other men of similar interests, came together again at Gresham College in Bishopsgate, London. After the restoration of the monarchy in 1660 these meetings became more regular and it was after hearing a lecture by Christopher Wren, then a professor of astronomy, in November of that year that the gathering decided to form a Society for promoting Physico-Mathematical Experimental Learning. It is the three-hundredth anniversary of this event which is being celebrated this year. Two years later, in 1662, King Charles II granted a royal charter of incorporation and in the following year he presented the mace which is one of the Society's most treasured possessions; it is of silver, richly gilt and before business commences at any meeting of the Council or of the Society it is placed in front of the President.

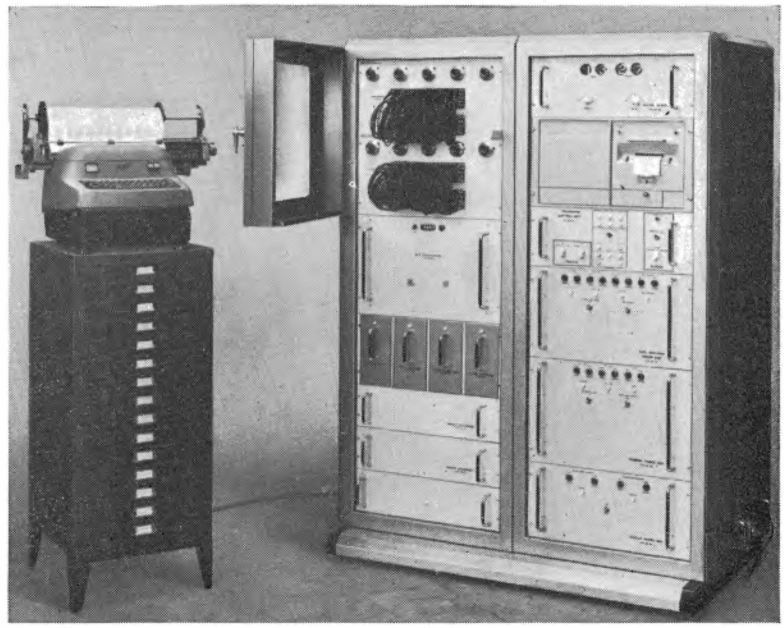
As is to be expected of any institution of so great an age, the Royal Society has had its share of ups and downs and at times, due perhaps to a shortage of funds, it has not been over fussy about those whom it has admitted to its Fellowship. Thus one may find in its Charter Book the names of such 'unscientific' personages as Dryden, Samuel Pepys, Edward Gibbon, Warren Hastings and even Lord Byron. Ranged alongside these, however, are the names of many of the all-time scientific 'greats' and among them are many of the men who, unwittingly perhaps, made our own particular branch of science possible; names which readily spring to mind in this context are those of Newton, Clerk Maxwell, Lord Kelvin, J. J. Thomson and Lord Rutherford. In more modern times the Society has, of course, been very particular indeed in admitting persons to the Fellowship. The present number of Fellows is about 600 plus 63 Foreign Members and, despite the vast increase in the number of men devoted to the pursuit of science, only 25 persons are admitted to the Fellowship at each annual election and to be so elected may be considered as, perhaps, the greatest of scientific honours.

The Royal Society is not, however, merely an historic institution or a place where learned papers can be read and eventually published. It carries out a great deal of practical work. High in importance in its activities is the responsibility of its President and Council for the management of the scientific work of the National Physical Laboratory which, incidentally, the Society took a leading part in founding; it also administers, on behalf of the government a number of annual grants voted by Parliament in aid of scientific investigations and scientific congresses etc. Another national responsibility of the Society lies in the field of international relations in science where it does much valuable work in connexion with the International Council of Scieintific Unions and UNESCO and in this sphere one cannot neglect to mention the very fine work which the Society has performed in connexion with the recent International Geophysical Year.

This year then is a momentous occasion for the world of science and it is with great pleasure and due humility that we tender our congratulations and good wishes for its continued progress to the Royal Society.

A HIGH SPEED DIGITAL DATA LOGGING SYSTEM

By H. Fuchs*, B.Sc, Ph.D.,
and J. R. Wastell*, B.Sc.



This article describes a high speed digital data logging system which has been designed to monitor the outputs from a large number of electric transducers, typically thermocouples, strain gauges, resistance bulb thermometers, pressure transducers, etc.

The modules from which such a system is built up are described and their performance discussed. In particular, it outlines the design of a scanning unit capable of switching microvolt signals at fairly high speeds, the design of a data amplifier suitable for discontinuous operation and having extremely low noise and drift and a high speed analogue to digital convertor with an accuracy of 0·1 per cent and a decimal output.

The system, which is intended for 24 hour operation, incorporates its own automatic checks and is capable of detecting departures from normal operating conditions and producing a complete statement of the conditions at regular intervals of time.

(Voir page 526 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 531)

THIS article describes a high speed digital data logging system which has been designed for use with electric transducers, typically covering the parameters temperature, pressure, flow rate, etc., on industrial and nuclear plant.

The equipment scans the voltage outputs of the transducers at a high speed. The output from each parameter is examined by the equipment to ensure that it has not departed from pre-set working limits.

If such a departure has occurred, the equipment is immediately energized to produce an output, in digital form, on a printer or tape punch. This output comprises the identification of the parameter, its value and a symbol indicating whether the departure is an increase or decrease in the parameter value. At regular pre-determined intervals of time, and also on demand by the operator, the equipment will produce a complete digital log on an electric typewriter, giving the identification and parameter value of every point scanned.

The equipment is designed on a modular basis making it extremely easy both to design a system to meet a particular specification and to change the specification at some later date.

The equipment is completely automatic in operation and performs checks upon itself once every scan.

Description of System

A data logging system may comprise any or all of the following sub-assemblies, each of which is constructed as a removable sub-assembly and the whole system being in standard 19in equipment racks.

The units comprising the data logging system are:—

(a) Input scanner.
(b) Scanner drive.
(c) Data amplifier.
(d) Linearization and range change.
(e) Alarm detector.
(f) Alarm level setting unit.
(g) Analogue to digital convertor.
(h) Digital clock.
(i) Programme control unit.
(j) Digital recorders.
(k) Digital store and display.
(l) Power supplies.

A typical information flow block diagram is shown in Fig. 1.

Transducer outputs are scanned by the input scanner which is itself driven by the scanner drive at pre-determined speeds compatible with the output recorder being used.

The output from the input scanner is amplified in the data amplifier and then is supplied to the analogue to digital convertor which converts the information into digital form suitable for driving the digital recorders and the digital store and display.

The output of the data amplifier also operates the alarm detector which is itself supplied by information from the alarm level setting unit selected by the input scanner.

If the alarm detector detects that the parameter voltage has departed from the pre-set limits, an alarm relay is de-energized which commands the programme control unit

The above photograph shows a typical 100 point data logger

* Blackburn Electronics Ltd

to supply the appropriate output recorders with information.

A digital clock supplies the time to the output recorders and also generates the logging signals at pre-selected intervals of time, selection being performed by decade switches on the front panel of the clock.

Description of Sub-Assemblies

Input Scanning Unit

The scanning unit switches each transducer output in turn to the data amplifier and in addition selects an identification code and an alarm level appropriate to each point.

Scanning Drive Unit

A hard valve clock pulse generator comprising a bistable circuit driven by a blocking oscillator is used to switch each relay in turn. The basic circuit diagram is shown in Fig. 2 and it will be seen that the system operates by switching the current alternately through valve 1 and valve 2 and selecting the appropriate relay to be energized by the cycling contacts.

Since no mechanical part of the scanning system runs at the scanning speed, a very long life is ensured, a typical figure for a 100 point scanning unit being 300 000 hours continuous running.

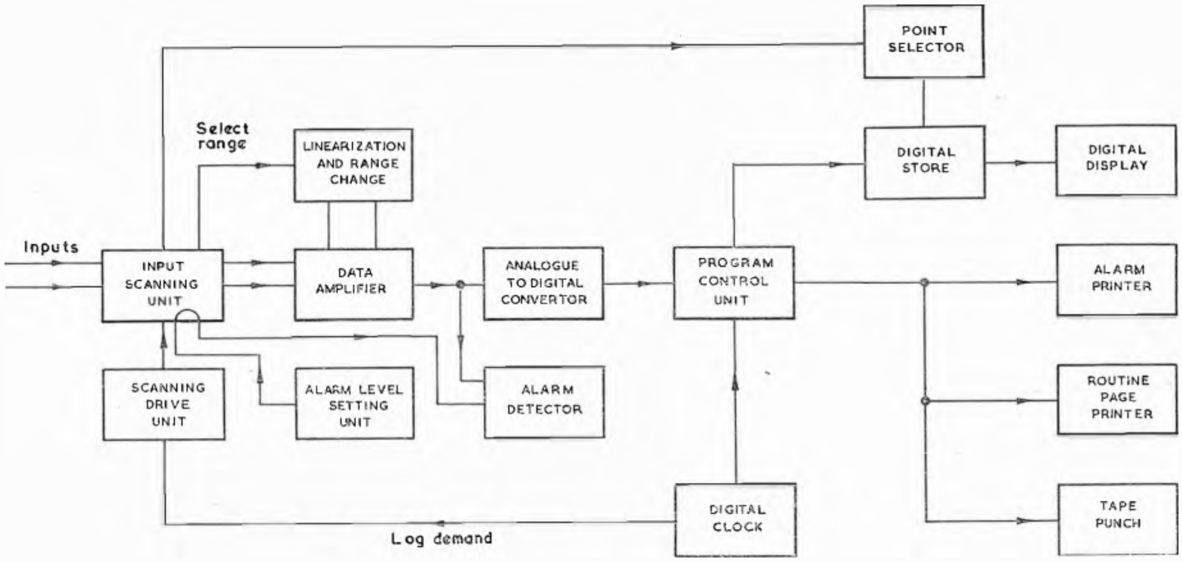


Fig. 1. A typical data logging system

The basic switching element employed is an hermetically sealed relay with gold plated contacts.

The scanning unit is capable of running at a speed of 50 points per second when no printing is being demanded, i.e. when alarm scanning. When printing occurs, the scanning speed is slowed down to a speed compatible with the particular printer being used.

The scanning unit is engineered in blocks of ten channels and any number of such channels may be incorporated. Any block of ten channels may be switched out by the operation of a toggle switch. Under these circumstances, the scanning unit omits these ten channels which are then available for preventive maintenance.

The toggle switches may be replaced by relay contacts in which case a programmed scan may be obtained.

The functions the scanning relays form are

(a) Input selection.
(b) Alarm voltage selection.
(c) Identifying code.
(d) Cycling.

Functions (a) and (b) are performed as stated above by gold plated contacts since analogue voltages are involved.

Functions (c) and (d) are switching operations *per se* and are performed adequately well by cheaper contact materials.

Each point consequently comprises two relays, one a sealed relay with gold plated contacts and the other a standard relay.

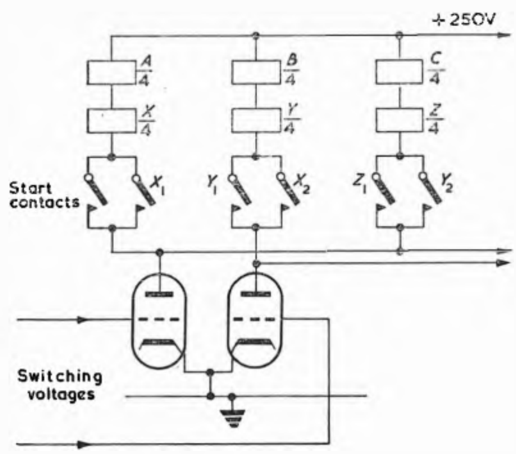


Fig. 2. The basic scanning circuit

Data Amplifier

Each input is switched in turn to the data amplifier which is used to bring the transducer output voltage to a level suitable for driving the analogue to digital convertor.

The maximum output voltage from the transducers will typically be 10mV full-scale and a gain of 1 000 in the data amplifier results in an output of about 10V. This is compatible with the input of the analogue to digital convertor.

Drift in the amplifier is kept to less than 10μV referred to the input by correcting for the drift between each switching operation, as shown in Fig. 3.

The amplifier has an in-phase rejection ratio in excess of 30 000 :1, the inputs floating, in order to cope with any pick up voltages likely to be present to the transducer leads.

The data amplifier comprises three units each having an internal gain of 30 000 and a high degree of feedback.

The first stage has an effective gain of -1 and the second stage sums the output from this amplifier, which is fed by one of the two input lines, and the output of the other input line. This amplifier has a gain of approximately -30.

Summing at the input of the second amplifier provides the high in-phase rejection ratio and the floating input.

A third stage with a gain in the region of -30 includes the drift correction and the linearization circuits.

The drift correction is achieved by charging a capacitor

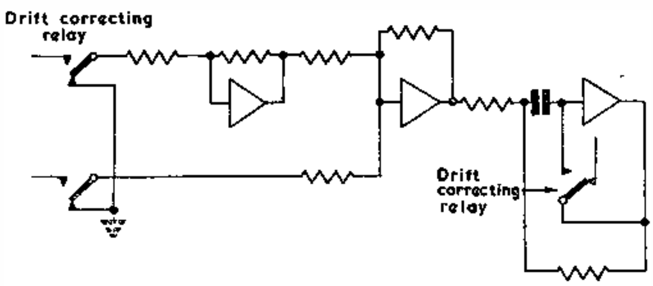


Fig. 3. The data amplifier

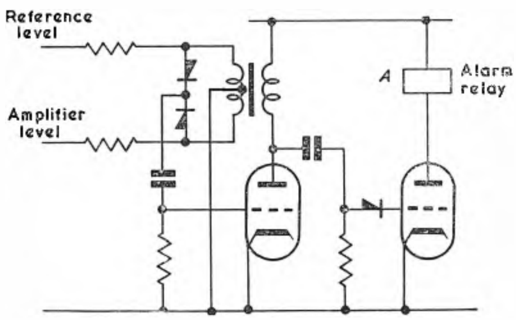


Fig. 4. Basic alarm detector circuit

at the input of the third amplifier with the total drift voltage between each measurement by short-circuiting the input of the data amplifier and connecting the output to a drift correcting capacitor.

Linearization and Range Change Unit

The total gain of the data amplifier is set by precision wire-wound resistors, setting the feedback ratios.

Should linearization be required, a diode function generator is incorporated in the feedback loop of the last amplifier.

The break point and fractional gain change may thus be set empirically to achieve the best approximation to a straight line. A typical example of this is the linearization of a chromel/alumel thermocouple output in the range 0 to 1 000°C where a single kink occurs in the characteristics at about 400°C. This kink may be taken out with a single diode.

Should several transducers be scanned with different scale factors, the appropriate gain is selected by the input scanner, by switching the appropriate gain setting resistors into the data amplifier.

Should the thermocouple transducers require cold junction compensation, a single cold junction in the input of the data amplifier is provided which is switched in turn to each thermocouple.

Alarm Level Setting Unit

Alarm levels may be set for each point, either by ten-turn helical potentiometers providing infinitely variable levels, or by a precision potentiometric chain, providing discrete levels.

In either case, the alarm levels are brought to a patch panel which permits any input point to be allocated any alarm level. Patching is performed remotely from the logging equipment and the pre-programmed board may be then connected to the logger in one simple operation permitting the change of alarm levels without interfering with the logging scan.

A transistor reference power supply provides the stable voltage for supplying the alarm level setting potentiometers.

This unit is enclosed in an isothermal oven and has a stability of 0·05 per cent under all conditions.

Alarm Detector (Fig. 4)

The alarm detector compares the output voltage from

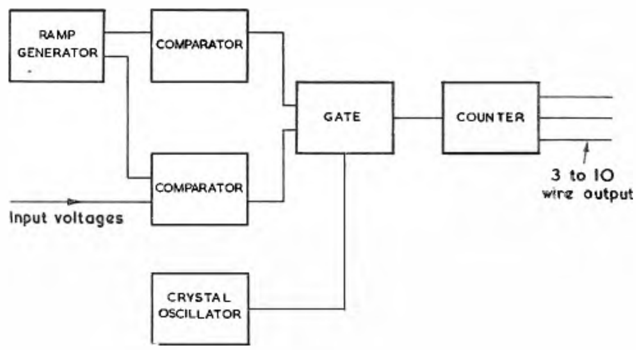


Fig. 5. The analogue to digital convertor

the data amplifier with that from the level setting potentiometer appropriate to that point. Should the output voltage exceed the set level voltage, a relay is de-energized.

A diode bridge in the feedback loop of a blocking oscillator, is used as the voltage sensing element, eliminating dependence upon diode characteristics.

The alarm detector has a backlash and drift of 10μV referred to the input of the data amplifier.

Analogue to Digital Convertor (Fig. 5)

The analogue to digital convertor converts the output from the data amplifier to a form suitable for driving the output printers.

The unit represents the input voltage as a three or four digit decimal number.

A linear ramp voltage is generated and compared with earth potential in one multiar and with the incoming potential in a second multiar comparator.

The output pulse from the first multiar opens a gate while the output pulse from the second multiar closes the gate.

Pulses from a 200kc/s crystal controlled oscillator pass, during the time the gate is open to a three Trochotron or four Trochotron counter. At the completion of the count, therefore, the number of pulses counted are a measure of the size of the input potential.

The output from the convertor is an 80V negative step on three of thirty lines or four of forty lines. The conversion time for three decimal digits is 5msec and for four decimal digits is 50msec.

The total errors in the digital output are the sum of the errors introduced by the data amplifier and the analogue to digital convertor.

Total errors introduced by the data amplifier are 0·2 per cent of full-scale and those introduced by the analogue to digital convertor are 0·1 per cent of full-scale.

The total r.m.s. error is hence 0·25 per cent.

Programme Control Unit

This unit routes all the information to the output recorders and display equipment and also contains the logical circuits which control the sequencial operation of the printers under alarm conditions.

The digital output from the analogue to digital convertor, the digital clock and point identification information from the input scanner is routed to the printers via this unit which selects which information is to be printed. In the case of the strip printer, all information is routed out in parallel but in the case of the typewriter, the information is serialized and fed to the printers one digit at a time.

The programme control unit also carries a centralized alarm panel comprising the following alarm lamps:—

Power supplies.
High parameter value.
Low parameter value.
Open-circuit.
Short-circuit.
System check.
Scanner fault.
Typewriter fault.

Controls for the scanner and printers are also mounted on the front panel of this unit, giving the facility of scanning continuously looking for alarms, scanning at intervals selected on the clock, or performing one scan upon pressing the 'scan demand' button.

Each printer may also be used as a logging printer, i.e. printing out all information scanned at regular intervals of time or as an alarm printer, i.e. printing out only identification and value of alarmed parameters.

Digital Clock

The digital clock generates the time in hours and minutes for driving the output printers and also for initiating logging at regular intervals. It includes a four digit Nodistron display which indicates the time and may be used for setting up the clock initially.

The unit is synchronized from the 50c/s mains by means of a synchronous timing unit which operates a microswitch once per minute. These pulses drive the 'minutes' uniselector which in turn drives an 'hours' uniselector.

Two decade switches permit logging intervals to be set as follows:—

Minutes—5 10 15 20 30 60
Hours —2 4 8 12 24

Digital Store and Display

Any number of digital stores and displays may be incorporated with the equipment.

The functions of the store and display are to enable input parameters to be monitored at the high scanning speed without having to slow down to a speed compatible with conventional output printers. The store in operation, is supplied, once per scan, with the digitized parameter value from the selected input.

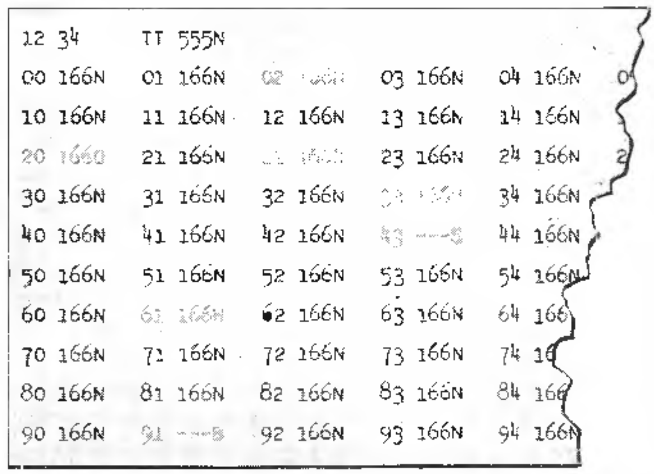
In the intermediate period, the information is stored and displayed continuously. When the scanning unit again returns to the selected input the store is reset and supplied with the new digitized value of the parameter.

The store comprises a number of printed circuit boards (being one decimal digit) each having ten cold-cathode trigger tubes. The appropriate trigger tube is struck by the output pulse from the analogue to digital convertor and the change in potential on its anode then illuminates the appropriate numeral in the Nodistron display tube.

Selection of the required input to be displayed is performed by decade rotary switches on the display unit and only involves switching a single line. This is achieved by driving all displays in parallel at all times but only energizing the appropriate store at the instant in time at which the input is connected to the selected parameter.

Output Recorders

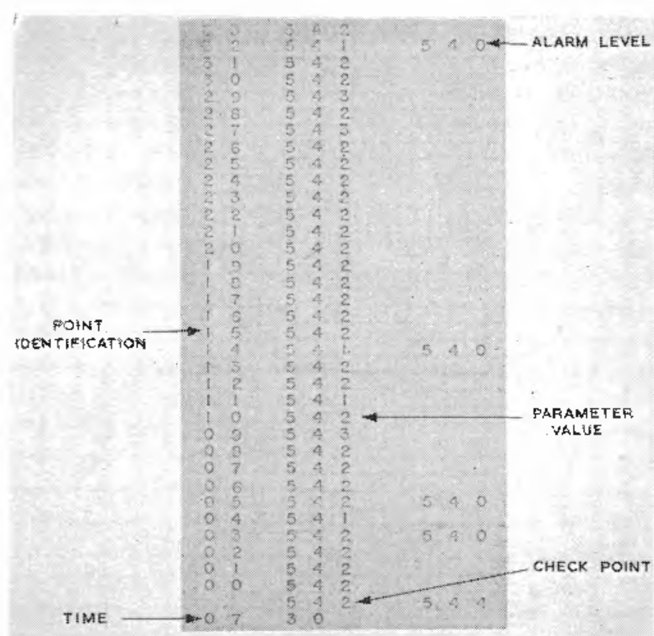
A number of printers may be used with the equipment ranging from a high speed strip printer, the Hewlett Packard 561B, capable of printing five, eleven digit, lines



12 34	TT 555N			
00 166N	01 166N	02 [illegible]	03 166N	04 166N
10 166N	11 166N	12 166N	13 166N	14 166N
20 166[illegible]	21 166N	[illegible]	23 166N	24 166N
30 166N	31 166N	32 166N	[illegible]	34 166N
40 166N	41 166N	42 166N	43 ---S	44 166N
50 166N	51 166N	52 166N	53 166N	54 166N
60 166N	61 166N	62 166N	63 166N	64 166
70 166N	71 166N	72 166N	73 166N	74 1
80 166N	81 166N	82 166N	83 166N	84 166
90 166N	91 ---S	92 166N	93 166N	94 166N

Fig. 6. A section of a typical log sheet obtained from an electric typewriter
On the original values in 'alarm' are printed in red and are reproduced here in grey

Fig. 7. A section of a typical log sheet obtained from a strip printer



per second; a conventional electric typewriter, capable of printing at about 10 digits per second; or a paper tape punch, the Creed Mark V, capable of printing 32 characters per second.

Under all output recording conditions, the information recorded normally comprises

Point identification—3 decimal digits or characters.
Space —1 decimal digit or character.
Parameter Value --3 or 4 decimal digits or characters.
High/Low/Normal/Faulty Indication—1 digit or character.

Time as four decimal digits and the check point for the equipment, are usually printed at the beginning of each scan.

On the typewriter, faulty points or alarmed points are printed in red.

A typical log sheet obtained from an electric typewriter is shown in Fig. 6 and a record from a strip printer in Fig. 7.

Power Supplies

The equipment operates from the 110-240V 40-60c/s single phase mains supply.

Stabilized h.t. is provided for the reference level unit at 50V 200 mA with a stability of 0·05 per cent.

The analogue to digital convertor has its own built-in stabilized high tension supplies at ±250V.

Low voltage supplies for relay and stepping switch operation are provided at 50V 5A and valve heaters operate at 6·3V a.c.

Stabilized h.t. is provided for the digital stores and displays at 350V and 250V 150mA.

System Checks

A number of checks on the operation of the equipment are performed while it is functioning.

At the commencement of each scan, a known voltage obtained from a standard reference cell is fed to the input and compared with an upper and lower limit in the alarm detector. If the output does not fall within the prescribed limits, an alarm condition results.

If an open- or short-circuited input transducer lead exists, the equipment detects this by noting that the output of the data amplifier is either excessively negative or excessively large. This detection is performed in two further alarm detectors.

Under these conditions the identity of the transducer is printed.

Blanks are printed in place of the parameter value and the appropriate symbol indicating a broken or short-circuited transducer is also printed.

Should the scanning unit cease to operate, due to any fault whatsoever an alarm condition is immediately detected and the faulty relay may be identified, by demanding a scan, the printed record so obtained stopping at the faulty point.

Should any power supply fail, an alarm condition is indicated by the de-energization of a relay which extinguishes the appropriate lamp.

All alarm conditions de-energize their appropriate alarm relay and extinguish the appropriate alarm lamp on a centralized warning panel.

In addition, all alarm conditions sound an alarm bell.

Acknowledgments

The authors wish to acknowledge the assistance of the staff of Blackburn Electronics Limited, Design Laboratory, for their work and the Directors of Blackburn Electronics Limited for permission to publish this article.

A Car Driving Trainer

A basic car driving trainer—the Sim-L-Car—which will enable drivers to become completely at ease in a motor car and to get the 'feel' of driving before venturing on the road, has been developed by General Precision Systems Limited, Aylesbury, Buckinghamshire.

The equipment is unique in that any type of car can be used and the characteristics accurately translated on to an 8ft × 4ft screen in front of the car. The screen shows a typical roadway complete with road signs, side roads, telegraph poles, hedges and trees. The driver, with an instructor beside him, can 'drive' anywhere he chooses within the limits of the shadowgraph model at any speed between zero and 60 mile/h.

The Sim-L-Car comprises a shadowgraph linked to the car's controls. A stationary point of light, level with the driver's eye, throws the picture on the screen from a disk on which the roadway model is built. The disk moves about the light source in response to the driver's use of the controls. This results in a highly realistic moving scene.

The noise of the car is electronically simulated and a control panel, held by the instructor, reflects the pupil's use of the gears, clutch, brake, etc. This means that where advanced driving techniques, such as double de-clutching, are being taught the instructor can see that they are carried out correctly. Attached to his panel is a speedometer showing the simulated road speed in miles per hour.

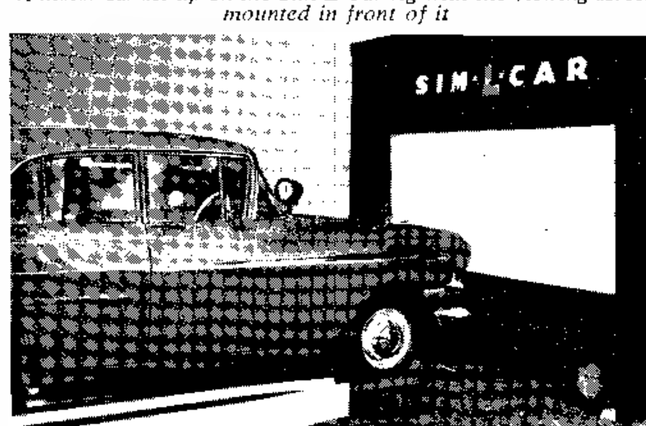
Should the driver touch the side of the road a buzzer sounds in the car. Additional refinements include a reaction timer measuring in tenths of seconds the time from the instructor signalling an emergency stop and the pupil depressing the brake pedal.

To set up the Sim-L-Car, the car to be used is driven on to a ramp and the front wheels guided into a pair of swivel units. These allow the steering wheel freedom of movement throughout its full range. The normal steering 'feel' of the car to be used can be obtained through adjustment of the units. The off-side unit is connected to the shadowgraph and by it the position of the car relative to the picture on the screen is dictated.

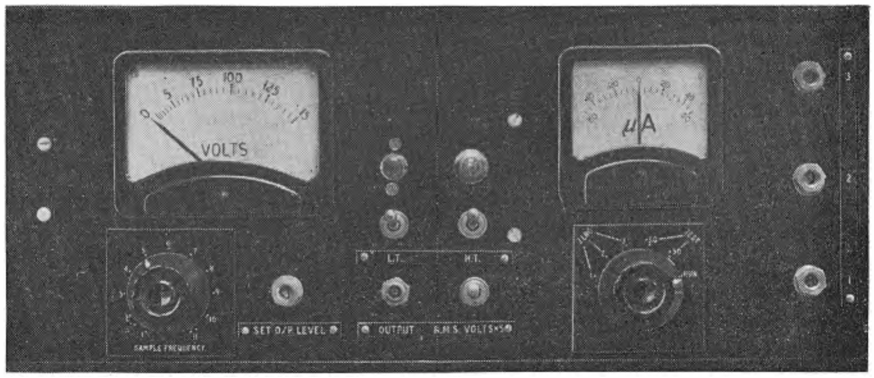
The normal foot pedals are used but behind them are slipped three separate controls mounted on a panel designed for the particular car. When the driver uses any foot control, the coincident pedal is actuated and this again affects the movement of the shadowgraph.

For the driving school the advantages of a Sim-L-Car are many. It can be used irrespective of weather conditions such as ice, snow or fog and during the long winter evenings. Furthermore there is no wear and tear on the car, no insurance, no maintenance and no running costs except for a negligible amount of electricity.

A motor car set up on the Sim-L-Car rig with the viewing screen mounted in front of it



A Low Frequency Noise Generator



By N. T. Slater*, B.Sc. (Hons.)

By means of a sampling technique 90 per cent of the noise power present in a high frequency band is converted into low frequency noise. This enables a high output level at low frequency to be obtained without the use of a high gain d.c. amplifier and its inherent drift characteristic. Simple and effective control keeps the source output constant and a method for accurate measurement of the final output power is described. The amplitude distribution of the output is equal to that of the primary noise source.

The generator operates entirely by electronic means using no moving mechanical parts.

(Voir page 526 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 531)

A NOISE generator was required for use with a 'LACE' analogue computer[1] to simulate the effect of atmospheric turbulence in aircraft problems. The required frequency range of the power spectrum was 0·01 to 20c/s, with a Gaussian amplitude distribution and adequate output voltage to provide forcing for the analogue equations.

Generators have been built elsewhere[2] using relays or commutator-bar elements; the one described operates entirely by electronic means and uses no moving mechanical parts. A simple and effective method of control keeps the source output constant, and means for continuous measurement of the final output is described.

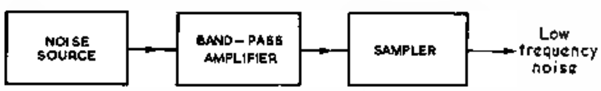


Fig. 1. Basic system

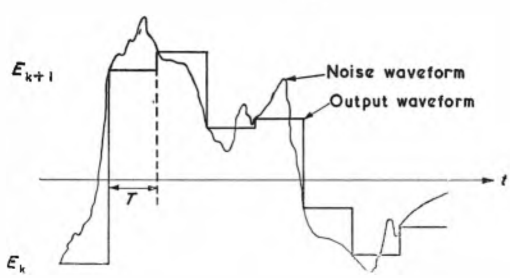


Fig. 2. The waveforms

Theory of Operation

All available noise sources have only a small electrical output, so that high amplification is necessary to give an acceptable output level. In this application the spectrum must also contain frequencies down to 0·01c/s. If a high-gain d.c. amplifier was used, the inherent drift of such an amplifier would affect the shape of the low-frequency end of the power spectrum. This difficulty can be overcome by amplifying at high frequency and transferring the power to the required frequency range by a sampling process.

The essentials of the system are shown in Fig. 1 and the output waveforms in Fig. 2. The amplitude distribution of the output will be the same as that of the primary noise.

It is necessary to calculate:

(1) The r.m.s. value of the output power.

(2) The shape of the output power spectrum.

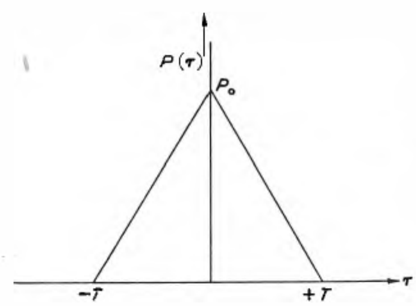


Fig. 3. Autocorrelation function

If it is assumed that successive samples E_k, E_{k+1} are not correlated (i.e. pure white noise is being sampled) then the auto-correlation function has the magnitude and shape indicated in Fig. 3. From this the power spectrum $P(\omega)$ can be calculated as:

$$P(\omega) = \frac{P_o T \sin^2 \pi f T}{(\pi f T)^2}$$

where

T = sampling period

f = frequency in question

P_o = total power of the noise being sampled.

The shape of the power spectrum is indicated in Fig. 4. From this expression the following properties can be obtained:

(1) All frequencies are present except $f = 1/T$, $1/2T$, $1/3T, \ldots$

* *English Electric Aviation Ltd.*

The above photograph shows the control panel

(2) The total power in the output spectrum is equal to the power (P_o) in the spectrum being sampled.

(3) The power in any small band of frequencies near 0c/s is dependent on the sampling period T and the total power P_o in the noise waveform.

(4) It can be shown that 90 per cent of the total noise power lies in the frequency region from $f=0$ to $f=1/T$.

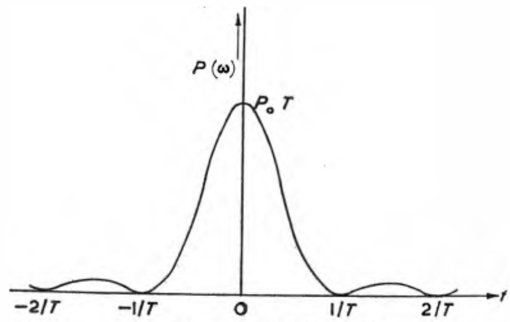


Fig. 4. The power spectrum

For such a diode:

$$i^2 = 3{\cdot}18\times10^{-19}\, I.\ \Delta f.$$

Where i = r.m.s. fluctuation (A).

I = mean diode current (A).

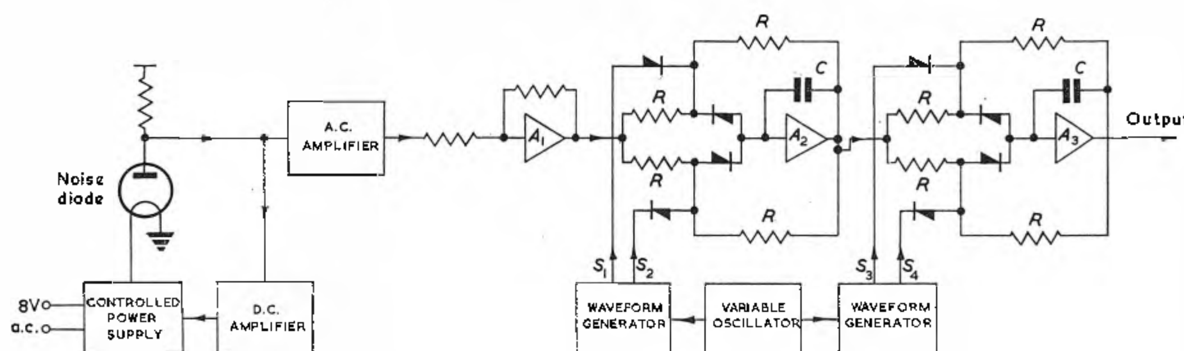
Δf = frequency band of noise.

If $\Delta f = 6\,000$c/s, $I=5$ma (CV 2 171)

then $i = 3\times10^{-19}$A.

If the noise voltage is developed across an anode resistor of 33kΩ then the source output is 100μV r.m.s. The a.c. amplifier thus requires a gain of 100dB in order to give an output of 10V r.m.s.

The d.c. voltage drop across the anode resistor gives a measure of the noise output, since this is proportional to the mean diode current. This signal can be used to control the anode current by adjustment of filament current. Continuous monitoring of the mean anode voltage is provided to ensure that it does not fall below the value required for saturated emission.



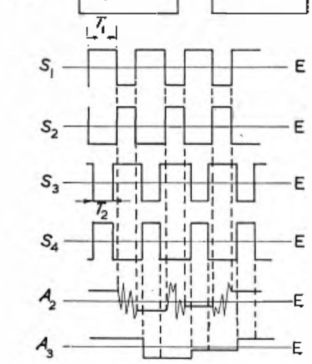


Fig. 5. Noise generator

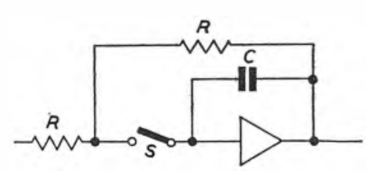


Fig. 6. The sample and hold system

Thus by means of the sampling process the bulk of the high frequency power has been transferred to frequencies near 0c/s.

In the above calculations it was assumed that pure white noise was being sampled, when in fact band-limited noise was used. The errors introduced by this can be shown to be small if the pass-band frequencies are high compared with the sampling frequency[3]. The pass-band frequencies in the generator were 2 to 8kc/s and sampling frequencies of 20 to 50c/s were used.

Noise Source

The problem of drift in d.c. amplifying stages has been overcome by the sampling technique, but in order to ensure a steady r.m.s. output, the output of the primary noise source must be accurately controlled.

Simple and effective control, together with comparatively high output, can be obtained by using a diode operated in temperature limited condition.

Diode Control System

A block diagram of the control system is included in Fig. 5. The d.c. anode potential is compared with a variable reference, and the error signal amplified by means of a long-tailed pair. The output from this amplifier controls the filament current through an OC16 transistor, the d.c. power being obtained from a simple transistorized supply fed from an 8V transformer tapping. Adjustment of the reference level enables the noise output to be varied over a small range within the saturated region of the diode.

It was found desirable to protect the filament from excessive current due to failure of the d.c. amplifier. Since the base driving voltage for the OC16 was about

$-4V$, effective protection was provided by connecting a 6V Zener diode between base and earth. This allows the base to move between 0 and $-6V$ only, holding the filament current to a safe value in the event of a large negative output from the amplifier, and holding the base at earth by the normal diode action if the amplifier output should try to rise to a large positive value.

The loop was designed to have low gain at the noise frequencies (2 to 8kc/s) so as not to affect the shape of the output spectrum, this property being provided to a large extent by the thermal lag in the filament.

The A.C. Amplifier

This amplifier has a feedback gain of about 100dB and a pass-band of 2 to 8kc/s. The input stage is cascode-coupled employing wire-wound resistors to keep unwanted noise signal to a minimum.

The gain of this amplifier is small at sampling frequencies (<50c/s) so that the negative feedback will not materially reduce the output impedance at these frequencies. To overcome this difficulty a d.c. amplifier of unity feedback gain is used as buffer between a.c. amplifier and sampler.

Sample and Hold System

Fig. 5 shows the method of sampling, together with associated waveforms. The operation of the circuit is identical with that of Fig. 6; the diode circuit performing the function of switch *S*. Whenever the switch opens the then present signal will be held by capacitor *C;* when the switch is closed the circuit is an ordinary invertor of unity gain, provided that the time-contsant *RC* is small compared with the period at the highest noise frequency (8kc/s). *C* must be large enough to prevent significant voltage change during the hold period due to leakage and grid current. Values of 10kΩ and 250pF were found to give satisfactory sampling.

The first switch opens for a period T_1 (say) and closes for a period $(T-T_1)$ where T is the period of the switching waveform. The second switch closes for a period T_2 and opens for $(T-T_2)$. If $T_1 > T_2$ the output will have the desired shape. This form of sampling prevents short transients produced at the instant of switching levels in the first switch being fed through to the output. These can die out in the interval $T_1-T_2/2$ before the second switch takes its sample.

This sampling system was found to give a clean output free from spikes or transients.

The amplifier A_1 has to operate at frequencies up to 8kc/s, and gives 50dB gain at this frequency. The voltage capability was ± 50, representing one machine unit on the computer giving a peak-to-r.m.s. value of 5:1 for 10V r.m.s. noise output. The peak signal of 50V was found to occur after relatively long time intervals and in use the r.m.s. value was very often amplified on the analogue computer at the expense of peak-to-r.m.s. ratio.

One difficulty with such a system is that when the switch *S* is open the amplifier is prone to 50c/s pick-up, since the 250pF capacitor presents a high impedance to signals at this frequency. To reduce pick-up to a minimum the switch diodes (EB91) were operated with d.c. heater supplies and the usual screening precautions were taken.

Noise Measurement

A thermocouple meter is used to measure the r.m.s. output value. In order to obtain a steady reading the thermal time-constant must be long compared with the lowest noise frequency being measured. This condition is necessary because the instantaneous value of random noise varies and it is the time averaged which is required. Since the meter pointer oscillates for inputs of less than 1c/s the above condition is not satisfied at the final output.

The r.m.s. value of the final output is equal to that of the band-limited noise, so the meter will read the correct value if placed before the sampler. It will now have a high-frequency input and will give a steady reading.

Because no continuous monitoring of the final output is provided, two d.c. test signals of ± 50V are provided to check gain and capability of the sampling system. This check is normally carried out each time the d.c. amplifiers are zeroed.

Conclusion

The variation in mean output level was less than 0·2V/h, this figure being entirely dependent on the d.c. invertors. If chopper-stabilized amplifiers were used the drift could be much reduced, enabling the generator to be used at lower frequencies. With the d.c. amplifiers used no appreciable signal above 0·01c/s was present at the output when the noise source was disconnected.

The output waveform of the generator has been analysed on a DEUCE digital computer, and value of output, amplitude distribution and shape of power spectrum were found to agree with the predicted values.

Although the output power spectrum is not flat over a wide range its shape is known and if necessary can be allowed for in an analogue computer problem.

Acknowledgment

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An Engine Telemetering System

The usual methods of obtaining information from moving parts on the variations of physical quantities (e.g. strain, temperature, relative displacement etc) require the use of devices such as slip-rings, flexible leads or intermittent contacts.

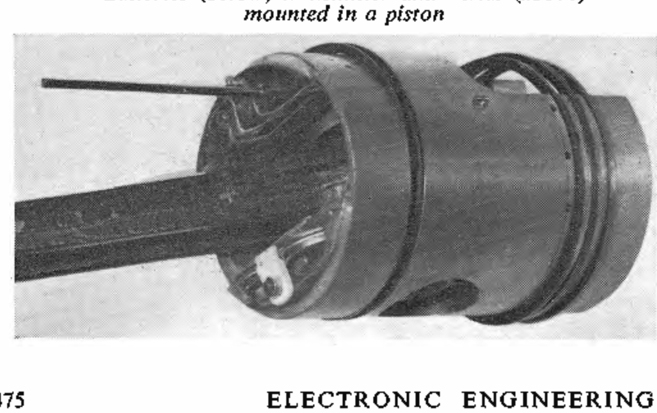
To avoid the disadvantages of such devices the British Internal Combustion Engine Research Association has developed a transistorized radio telemetering system using small but robust encapsulated units.

For instance, a transmitter complete with aerials and batteries may be mounted inside a piston reciprocating in a cylinder and the temperature at a selected point on the piston telemetered.

A carrier frequency of 2Mc/s is used and this is frequency-modulated by the transducer. At the receiver the frequency variations are converted into voltages suitable for display on an oscilloscope.

Similar equipment is used with conventional transducers for telemetering pressures, strains, displacements, and accelerations.

Batteries (below) transmitter and aerial (above) mounted in a piston



Systems Using Transistors and Transductors

By D. A. Ramsay*, B.Sc.

(PART 1)

Transistors are limited in voltage and power level and are sensitive to temperature variation. Transductors do not have these particular limitations and for some applications these two devices have complementary properties so that their combination provides advantages over the use of one or the other alone. In particular, a transistor control element can be incorporated with transductors to provide amplifiers which are not limited in output power or voltage. Also, highly efficient power amplifiers can be formed with circuits analogous to those of magnetic amplifiers, but in which the output transductor elements are replaced by switching transistors or by controlled silicon rectifiers; in these circuits transductors can be very suitable as control elements. Especially in conjunction with switching transistors, the useful properties of transductors can be exploited to provide small, robust and highly efficient systems.

(Voir page 526 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 531)

A TRANSDUCTOR is a device consisting of one or more ferromagnetic cores with windings which utilizes saturation phenomena in the magnetic circuit to control electrical power. A 'transductor element' is one of the cores with its windings which forms part of the transductor.

The most common use of transductors has been in magnetic amplifiers, which are electrical power (or signal) amplifiers in which amplification is produced by transductors. However, the non-linear behaviour of the magnetic core of a transductor element provides a number of useful effects besides amplification[1] and some of these effects are described below.

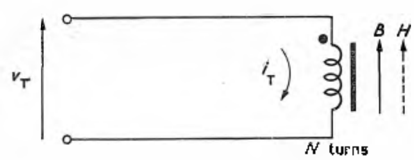


Fig. 1. The transductor element
A = effective cross-sectional area (cm²)
l = mean length of magnetic path (cm)

Transistors are compact, robust and reliable components which can operate efficiently down to low power levels. To retain these advantages in complete equipments there is a trend towards using transistors exclusively as the active elements in these equipments. For some functions, however, the temperature sensitivity of transistors and their limited voltage and (at present) power ratings can either require that other types of components be used or lead to undue circuit complexity if transistors alone are used. Transductors, which do not have these particular limitations, but which can have in common with transistors the advantages listed above, may then in some cases be combined with transistors to provide the best method of performing a particular function. Several ways are described below of combining transistors and transductors to provide advantages over the use of either one or the other type of component alone.

In particular, a transductor can incorporate a transistor control element to provide amplifiers which are not limited in output voltage or power. A design method for optimizing this combination is described in some detail.

B-H Loops

If the following approximations are assumed to be valid for a transductor element (Fig. 1):

(a) The flux density B and the magnetizing force H are uniform throughout the volume of the core
(b) There is no leakage flux
(c) The IR drop in the winding is zero

then the electrical and magnetic circuits of the transductor are related as follows:

$$v_T = 10^{-8} NA\,(dB/dt) \quad \ldots\ldots\ldots\ldots (1)$$

(It is convenient in this work to adopt the sign convention that a positive applied voltage induces a positive rate of change of flux density.)

$$B(t) - B(o) = (10^8/AN)\int_0^t v_T\,dt \quad \ldots\ldots\ldots\ldots (2)$$

$$i = ((10/4\pi)\ (l/N))\,H \quad \ldots\ldots\ldots\ldots (3)$$

$$H = ((4\pi/10)\ (N/l))\,i \quad \ldots\ldots\ldots\ldots (4)$$

If $v_T = v_{T(pk)} \sin \omega t$, then from equation (2):

$$B(t) = B_{(pk)} \sin(\omega t - \pi/2) + [B_{(pk)} + B(o)] \quad \ldots\ldots (5)$$

where:

$$B_{(pk)} = (10^8/AN)\ (v_{T(pk)}/\omega) \quad \ldots\ldots\ldots\ldots (6)$$

If the IR drop in the winding is not zero but is, say, a few per cent of $v_{T(pk)}$, then $B_{(pk)}$ will be a few per cent less than the value given in equation (6). Also the flux density waveform is then modified from that given in equation (5); the periodic component becoming not quite sinusoidal and the second term on the right-hand side having an exponential factor which decays to zero in about, say, ten to twenty cycles.

If the peak voltage $v_{T(pk)}$ exceeds a certain value then the peak induction $B_{(pk)}$ becomes sufficiently large for the core, some of the time, to be taken well into saturation. During such times, the winding current can be so large that assumption (c) becomes invalid and equations (1), (2), (5) and (6) cease, even approximately, to be true.

Summing up, if a sinusoidal voltage not exceeding a certain level is applied to a winding of a transductor element (other windings being open-circuits) then this impresses a nearly sinusoidal flux density in the core which, in the steady-state, swings symmetrically about zero. Corresponding to this periodic flux there is a periodic magnetizing force, also, in the steady-state, varying symmetrically about zero but in general severely distorted from the sinusoidal.

Unpolarized Magnetization

If a sinusoidal voltage is applied to a winding of a transductor element and the core induction B is plotted against the magnetizing force H, then in the steady-state, the resultant curve is a closed loop. Fig. 3(a) shows an observed family of loops corresponding to a set of values of applied voltage; the family shown is typical of 'rectangular loop' core materials (H.C.R. at 50c/s), i.e. materials for which B_r/B_s is nearly unity. When the peak value of the induc-

* *Applied Electronics Laboratories of The General Electric Co. Ltd, Stanmore, England.*

tion, $B_{(pk)}$, is very much less than the saturation value B_s, then to a first approximation the loops are ellipses. As $B_{(pk)}$ approaches B_s large peaks appear in the wave shape of the magnetizing force and hence in the magnetizing current.

If V_T, I_T denote the mean values (i.e. of the rectified waveforms) of v_T, i_T then the curve I_T/V_T (Fig. 2), termed the 'mean value magnetization curve' of the transductor element, has a shape which for a sinusoidal applied voltage at a given frequency is characteristic of the core material, and it is often used in transductor work. (Except for the sumbols B and H, capitals will subsequently be used to denote mean values of time variables.)

Transductor elements in magnetic amplifiers are usually operated with an applied alternating voltage which swings the flux density just short of the positive and negative saturation value, i.e. corresponding say to the second largest loop in Fig. 3(a) and to the knee of the magnetization curve in Fig. 2. The mean value of this alternating voltage will be referred to as the 'saturation voltage' of the element for the particular frequency and denoted V_s.

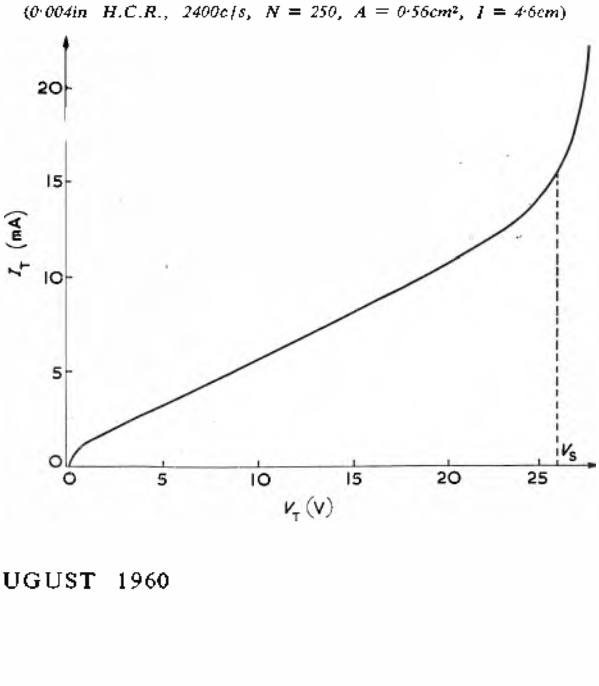
Polarized Magnetization

If a polarizing magnetizing force is impressed on a transductor element core (e.g. by passing a direct current through a second winding) then the *B-H* loops become asymmetrical with respect to the origin. Fig. 3(b) shows an observed family of steady-state asymmetrical *B-H* loops for a set of values of positive polarizing magnetizing force and a V_T corresponding to the saturation voltage (H.C.R. at 50c/s). As the polarizing force increases, then so does the fraction of a cycle during which the core is in saturation. With a very large polarizing force the core remains in saturation throughout the cycle. In this case assumption (c) is at all times invalid and the current through the a.c. winding is limited only by the leakage flux, the resistance of the winding and the impedance of the applied voltage source.

Fig. 3(c) shows a family of steady-state asymmetrical loops for the case where V_T is much less than the saturation voltage. To a first approximation all these loops are ellipses. As the polarizing force increases, the slope of the major axis of the *B-H* loop decreases, i.e. the incremental permeability of the core, and hence the inductance of the a.c. winding, decreases.

The area of a closed *B-H* loop is a measure of the total core loss per cycle. If, keeping $B_{(pk)}$ constant, the frequency of the applied alternating voltage is increased, then the eddy current loss in the core is increased and the loop

Fig. 2. Mean value magnetization curve
(0·004in H.C.R., 2400c/s, N = 250, A = 0·56cm², l = 4·6cm)



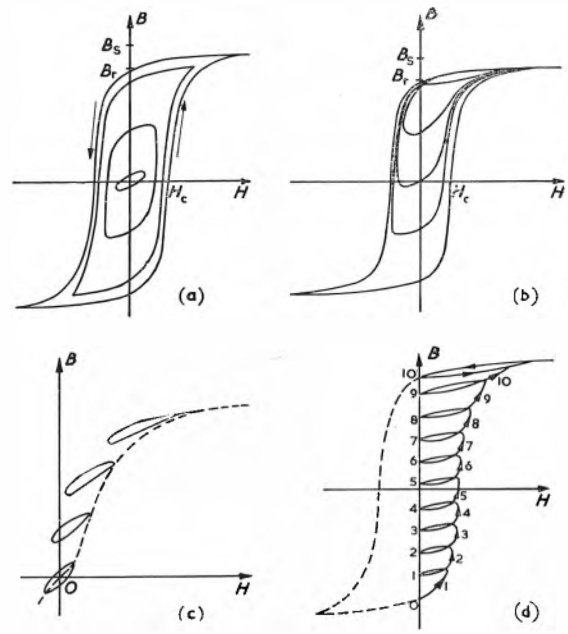


Fig. 3. The B - H loops
(a) Unpolarized (b) Polarized (c) Incremental (d) Pulsed (diagrammatic)

widens. Also if $B_{(pk)}$ and the frequency are both kept constant, but the flux waveform is distorted from the sinusoidal then the introduction of higher frequency components changes the eddy current loss and the area and shape of the loop. In some cases the distortion effect is small; for example severe distortion of the 50c/s alternating voltage applied to the transductor element when observing Figs. 3(a) and 3(b) changed *H*-values by no more than a few per cent.

In general, however, the shape of a steady-state *B-H* loop is a function of the intrinsic magnetic properties of the core material, its resistivity and the frequency spectrum of the *B* waveform or of the *H* waveform.

Pulsed Magnetization

So far closed steady-state *B-H* loops have been considered, corresponding to the application of an alternating voltage to a transductor element winding for an appreciable number of cycles. If a pulse of voltage is applied to the winding, then, provided the winding resistance can be neglected, equation (2) shows that the total change in flux density that results is:

$$\Delta B = (10^8/AN)\int_{t_2}^{t_1} v_T\,dt \quad \ldots\ldots\ldots\ldots\ldots (7)$$

where (t_1, t_2) is the pulse interval. If H (and hence the winding current i) is zero before and after the pulse then this increment ΔB is added to the initial flux density of the core. The initial flux density may have any value (depending on the previous history of the core magnetization) between $-B_r$ and $+B_r$ where B_r is the remanent flux density corresponding to H being reduced to zero from a very large value.

Fig. 3(d) is a diagram idealized to illustrate the *B-H* behaviour of a rectangular-loop material when a series of pulses of constant volt-seconds, each equivalent to say ΔB, is applied to a winding. The flux density initially corresponds to the point o on the *B*-axis where $B_0 = -B_r$. The first voltage pulse increases B to B_1 via curve 1, where $(B_1 - B_0) = \Delta B$. The second pulse increases B from B_1 to B_2 via curve 2, where again $(B_2 - B_1) = \Delta B$, and so on. In the Figure, ΔB is slightly less than $2B_r/9$ so that in this case nine pulses can be applied without taking the core into saturation. The first nine current pulses in the winding are therefore small, being limited to the magnetization current level. The tenth pulse, however, takes the core well into saturation so that a large current pulse flows, in this case being limited only by the winding resistance and the external impedance. The tenth pulse increases B only slightly, from B_9 to $B_{10} = B_r$. Further pulses correspond to a very shallow, very wide closed B-H loop which touches the B-axis at B_r and runs well into saturation.

If, in the above process, the volt-seconds value of each pulse is increased, then ΔB is increased and the number of pulses required to take the core from $-B_r$ to $+B_r$ is reduced.

The above is a much simplified account of pulse magnetization which is in fact a highly complex process (for example, see Melville[2]). In particular if the magnetizing force does drop to zero at the end of a pulse, a negative voltage will be induced across the winding which reduces somewhat the effective volt-seconds of the applied pulse, and by an amount which varies with succeeding pulses. Also, for equation (7) to be at least approximately true the series resistance in the winding circuit must be much less than the winding impedance. In this case the decay time-constant of the magnetizing force H is large and it is unlikely that H will have dropped to zero before the onset of the next pulse. Nevertheless, the essential features indicated by Fig. 3(d) can be usefully implemented in practice[3,4].

If a transductor winding is fed from a current source (i.e. a very high impedance source rather than the low impedance sources so far considered) there is a feature of the behaviour of the magnetic core which is of practical use. If a current pulse is applied to the winding, of a peak value not greater than that of a preceding current pulse, then the terminal flux density of the core tends to be little changed by the second pulse. The extent to which this is true depends on the material and design of the core[3]. Thus if a core is set to a datum flux density (e.g. $-B_r$) and a varying current is impressed on the winding, then the core flux density at any instant can be an approximate measure of the peak value of the current that has passed through the winding.

Useful Properties of a Magnetic Core

Switch

A winding on a core may present a very high impedance to an impressed voltage when the core is unsaturated; a very low impedance when the core is saturated.

Memory

The level of flux density in a core whose magnetizing force is zero is retained indefinitely. This level may have any value between $-B_r$ and $+B_r$, depending on the history of the voltages and currents applied to windings on the core prior to the magnetizing force being brought to zero.

Time and Frequency References

If t_1 is an instant when the core flux density of a transductor element is at $-B_s$ and t_2 is the instant when it is subsequently at $+B_s$, then from equation (2):

$$\int_{t_1}^{t_2} v_T\,dt = 2.10^{-8}\,AN\,B_s \qquad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

$$= \text{a constant for the transductor element.}$$

That is, a given element can absorb a fixed value of voltage-time integral or volt-seconds in going from negative saturation to positive saturation.

If v_T is constant over (t_1, t_2) then:

$$t_2 - t_1 = (2.10^{-8}\ AN\ B_s)\,.\,1/v_T \qquad \ldots\ldots (9)$$

i.e. the time interval between successive saturation instants (which can be sharply defined by the rapid rise in coil current) is inversely proportional to the voltage across the coil. Thus a known time interval can be generated by applying a known voltage across a transductor element winding.

The above process can be made recurrent so that a frequency proportional to an applied voltage is generated[3,4].

Voltage Reference

If a transductor element core is driven successively into positive and negative saturation by a sufficiently large alternating voltage at a constant frequency then equation (8) shows that the mean voltage generated across a winding is completely determined by the value of this frequency and the constants of the transductor element.

Variable Linear Inductance

A direct control current in a transductor element winding can vary the incremental permeability of the core (Fig. 3(c)). An a.c. winding on the core then forms an inductance whose value can be varied by the control current. Provided the voltage across the a.c. winding is small compared with the saturation voltage the incremental B-H loops are approximately, ellipses and hence the inductance is approximately linear.

Ferroresonance

A non-linear series or parallel resonant circuit can be formed with a transductor element and a capacitor. These circuits can have bistable characteristics which may be applied to pulse shaping and generation, logic operations, etc. They can also be used to generate oscillations which are sub-harmonics of the supply frequency.

Storm[1] classifies basic circuits using the properties described in the paragraphs above and lists their applications.

Basic Flux-Reset Magnetic Amplifier

Fig. 4(a) shows the basic form of a type of magnetic amplifier proposed by Ramey[5,6] in 1951. This type of amplifier is sometimes termed (for reasons which will appear below) a 'flux-reset' magnetic amplifier. An alternating supply voltage v_a is applied to a transductor element whose saturation voltage V_s equals V_a (the mean value of v_a) in series with two parallel branches: a load or output branch containing the load resistance R_L, and a control or 'reset' branch containing the control voltage v_c. For effective operation this type of magnetic amplifier requires cores for which B_r/B_s is nearly unity, ideally having B-H loops of the form shown in Fig. 4(c). (Compare Fig. 3(b)). To simplify the explanation of the behaviour of this circuit, v_c will first be given the form of a half-wave rectified sinusoidal voltage of the same frequency as v_a and independently variable in amplitude from zero to $v_{a(pk)}$, the peak value of v_a (Fig. 5(a) and 5(b)).

Suppose that at the start of a negative half-cycle of v_a, $\omega t = 0$ and the core has the initial conditions $B = B_s$, $H = 0$, i.e. the core operating point is at 1 on the B-H loop of Fig. 4(c). Over the following negative half-cycle of v_a, $\omega t = 0$ to $\omega t = \pi$, rectifier 1 is open, rectifier 2 conducts and the negative supply voltage diminished by the opposing control voltage v_c is impressed across the transductor element winding. This causes B to reduce, or 'reset', from B_s to B_M say, where, from equation (2):

$$B_s - B_M = (2.10^8/\omega AN)\,(v_{a(pk)} - v_{c(pk)}) \;\ldots\ldots (10)$$

Over this control or resetting half-cycle of the supply the core operating point in the B-H plane moves over the path 1-2-3-4 (Fig. 4(c)). During this half-cycle the current flowing in the circuit is limited to the very low magnetization current level i_s; this resetting current flows in opposition to v_c so that a small amount of power is returned to the control element supplying v_c.

Over the following positive half-cycle of v_s, $\omega t = \pi$ to 2π, rectifier 2 is open, rectifier 1 conducts and v_a is impressed across the winding in series with R_L. In practice, $R_L i_s \ll$

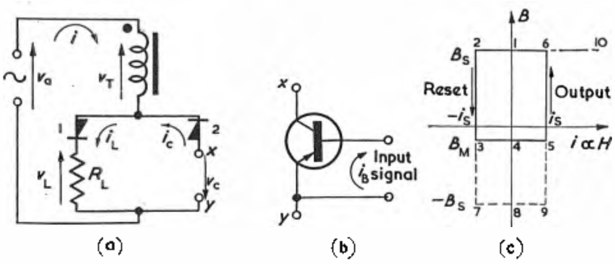


Fig. 4. Basic flux-reset magnetic amplifier
(a) with voltage source control element (b) transistor control element (c) idealized core B-H loop

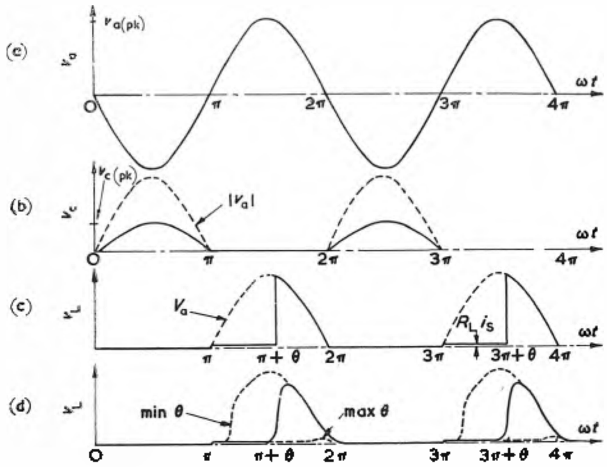


Fig. 5. Waveforms
(a) supply voltage (b) control voltage (c) load voltage: ideal core (d) load voltage: H.C.R. Core

$v_{a(pk)}$, that is the unsaturated impedance of the transductor element is made much greater than the load impedance. Thus so long as the core is unsaturated the supply voltage appears almost entirely across the winding. This positive voltage causes B to rise from B_M to reach B_S at some instant during the positive supply half-cycle, say at $\omega t = \pi + \theta$, where from equations (2) and (10):

$$\cos\theta = (B_M/B_S) = 2\,(v_{c(pk)}/v_{a(pk)}) - 1 \quad \ldots. \quad (11)$$

Over $\omega t = \pi$ to $\omega t + \theta$ the core operating point moves over the path 4-5-6 (Fig. 4(c)).

At $\omega t = \pi + \theta$, the 'firing angle', the core saturates, the winding impedance becomes almost nil and v_a is abruptly transferred from being almost entirely across the winding to being almost entirely across the load R_L. The supply current i jumps from the value i_s to the much larger value:

$$i_L = (v_s/R_L)$$

This large value of current through the winding holds the core in saturation until it drops to zero at the end of the half-cycle. Over $\omega t = \pi + \theta$ to $\omega t = 2\pi$ the core operating point moves over the path 6-10-1.

The output voltage waveform v_L for this circuit with an idealized core is shown in Fig. 5(c). Fig. 5(d) shows an actual v_L waveform for the core of H.C.R. for which B-H loops are shown in Fig. 3(b).

As $v_{c(pk)}$ is increased from zero to $v_{a(pk)}$, B_M increases from $-B_s$ to $+B_s$ (equation (10)) and θ decreases from 2π to π. If V_a, V_c and V_L are the mean values of the supply voltage, control voltage and load voltage respectively and i_s is assumed to be negligibly small then it can be shown from the foregoing equations that:

$$V_L = V_c \text{ for } 0 \leqslant V_c \leqslant \tfrac{1}{2} V_a \quad \ldots\ldots\ldots \quad (12)$$

Thus, provided mean values are taken as the measures of the output and input voltages, the system described above forms a linear voltage amplifier of unity gain. Fig. 6 shows the mean value voltage transfer characteristics of this amplifier (a) with idealized core (i_s not negligible); and (b) with H.C.R. core.

If the power gain G of the above amplifier is defined as the maximum ratio of the mean power dissipated in the load to the mean power dissipated in the control element, then it can be shown that:

$$G \simeq (V_a/R_L i_s) \quad \ldots\ldots\ldots\ldots\ldots \quad (13)$$

i.e. G is approximately equal to the ratio of maximum to

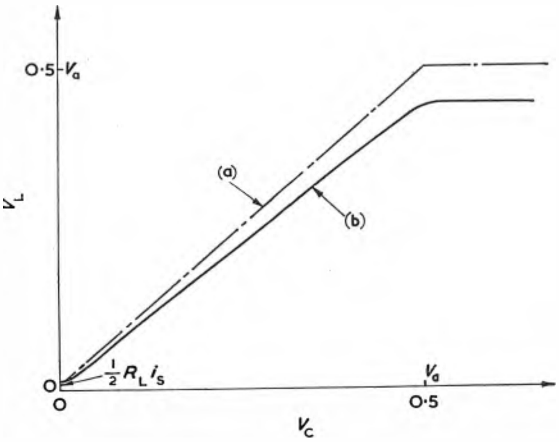


Fig. 6. Mean load voltage (V_L)/mean control voltage (V_c) curve
(a) Idealized core (b) H.C.R. core

minimum mean output voltage. In Fig. 4(c) the core power loss per unit volume is proportional to i_s. Thus G is approximately inversely proportional to the core loss per unit volume when this core loss is measured at the supply frequency and for saturation peak flux density.

Iron and copper losses are the major losses in the above amplifier. Efficiencies of above 80 per cent can be obtained, an upper limit being set by the core loss at the supply frequency.

An input voltage to the circuit of Fig. 4(a) must be present over a complete reset half-cycle before the firing angle of the succeeding output half-cycle has the steady-state value corresponding to this input. There is therefore a minimum time lag of one supply half-cycle between input and output. This amplifier is called a 'half-cycle response' or 'high-speed' magnetic amplifier to distinguish it from earlier types of magnetic amplifier whose response times are of the order of 10 cycles, rather than one cycle, of the supply.

If the amplitude of the control voltage is varied sinusoidally at signal frequency; the above amplifier responds to signal frequencies in the range zero to one quarter the supply frequency at the latter frequency the amplifier lag is 45° and its voltage gain is 3dB below its value at low frequency.

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(To be continued)

AN EDDY-CURRENT FLAW DETECTOR

By E. Ruff*, B.Sc.

Examination of airframe bolt-holes for cracks requires a quick non-destructive method. This article describes such an instrument using the effect on an inductance of a broken shorted turn, and gives some other related applications.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

FASTER speeds and greater utilization of aircraft are demanding even higher mechanical standards in the periodic examination of aircraft. The bolt holes of spar booms are now inspected for flaws and fatigue cracks to determine whether the boom has further 'safe' life. It was necessary to devise a method of doing this testing, such that an aircraft would be grounded for a minimum period.

Fig. 1. The instrument and accessories

As the personnel required to carry out the checking would be essentially unskilled electronically, an instrument was required which had a minimum number of controls and the simplest circuits consistent with reliability. The instrument to be considered measures 8in by 4½in by 3¾in and weighs 2½lb. It has three controls to be set by the operator, one of which selects the particular function for which the instrument is to be used. A small attache carrying case contains the instrument, together with a number of probes for different size bolt holes and spare batteries (see Fig. 1). A 4·5V torch battery was chosen for supplies as this was readily available internationally. The instrument is a comparator and can therefore only find flaws by comparison with a known good specimen.

Non-destructive Eddy-Current Testing

An eddy-current technique, using a probe to search the bolt hole, is employed, the electrical information gained being translated into terms of the magnitude of a defect in a bolt hole. The fundamentals of the technique can briefly be explained as follows. If the surface of an electrical conductor is irradiated with a sinusoidal alternating field, penetration of this field, the 'skin effect' obeys the law:

$$d = \left(\frac{\rho}{\mu_0 \mu_R \omega}\right)^{\frac{1}{2}} \text{ m.k.s. units}$$

where d = depth of penetration (metres)

ρ = resistivity of material (ohm-metres)

μ_0 = permeability of free space

μ_R = relative permeability of material

ω = angular velocity of alternating field (radians. sec^{-1}).

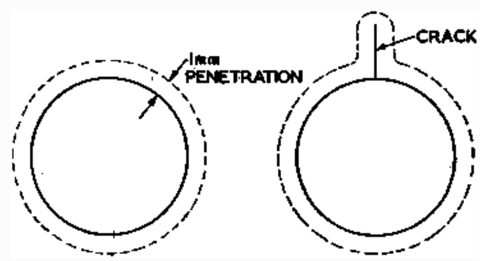

Fig. 2. Modification of eddy-current path length by crack or flaw

The total flux in a conductor is given by:

$$\Phi = \int_0^\infty B\,dx \qquad \text{where } B \text{ is the flux density}$$

$$= \int_0^\infty B_0 \exp(-x\sqrt{i}d)\,dx \qquad \sqrt{i} = \frac{(1+i)\sqrt{2}}{2}$$

$$= \int_0^\infty B_0 \exp\left[\frac{-x(1+i)\sqrt{2}d}{2}\right] dx$$

$$= \frac{B_0\sqrt{2}d}{1+i} = \frac{B_0 d(1-i)}{\sqrt{2}} \qquad |1-i| = \sqrt{2}$$

i.e. the total flux is the product of the surface flux density B_0 and the depth of penetration d. Consideration of the particular problem suggested that a value for d of 1mm would be ideal. The materials to be tested were all aluminium alloys and thus were very similar electrically, having the following nominal values for the relevant parameters:

$\rho \simeq 4{\cdot}5 \times 10^{-8}\Omega$ metres

$\mu_R \simeq 1$

The frequency of the alternating field must be approximately 6kc/s in order that d should be 1mm under these conditions.

Mode of Operation

A single 'search' coil sited in a probe irradiates the immediate surround of the bolt hole, as shown in Fig. 2. The 'search' coil is acting as the primary of a transformer

* *Bristol Aircraft Limited.*

with the periphery of the hole as a shorted-turn secondary. The effective eddy-current path length is shown for a good hole and the modified path due to a crack. This change in path length modifies the impedance of the shorted turn and thus the reflected impedance seen at the primary of the transformer, i.e. the 'search' coil. If the 'search' coil is incorporated as one arm of an a.c. version of a conventional Wheatstone bridge as in Fig. 3, this change of reflected impedance can be detected as a change of both phase and amplitude of the output of the bridge. In this case only the change of phase will be considered.

Circuit Details

Fig. 4 shows the layout of the instrument and the complete circuit is shown in Fig. 5. The output from the bridge is fed into a three-stage limiting amplifier. This consists of the three transistors $VT_{4,5,6}$, the first and third stages having d.c. feedback and stabilization. The diode MR_1 is included

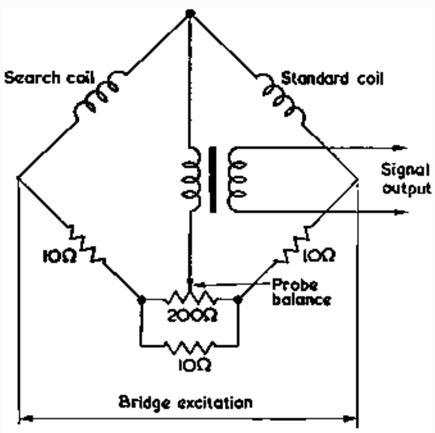

Fig. 3. Bridge network

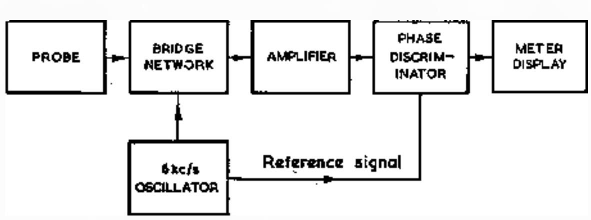

Fig. 4. The layout of the equipment

to improve the mark-to-space ratio of the resulting square-wave, being connected in the reverse sense to the base-emitter junction. This square wave is then fed to the collector of transistor VT_7 which acts as a modulator phase discriminator. A reference square-wave, in phase with the bridge excitation, switches the transistor 'on' and 'off'. Typical waveforms observed at the collector of VT_7 are as shown in Fig. 6. At (A) there is no signal output from the bridge and only the reference switching signal is present. At (B) the signal waveform, half the amplitude of the reference, leads the reference. It will be seen that if those waveforms are integrated about the zero level shown, which in this case is the centre of the waveform. (A) will give zero indication and (B) will give a positive indication. The instrument carries out this function using a d.c. microammeter. The zero level is controlled by applying a backing-off voltage to the earthy side of the meter. The meter integrates the waveform over a number of cycles and gives zero reading for the waveform at (A) and a positive reading for that at (B).

The oscillator, providing both the reference and bridge excitation, is a grounded emitter stage. An output of the correct phase is fed back from the collector of the transistor to the base. The amplitude of oscillation is controlled, over a limited range of loading conditions, by rectifying part of the signal from the transformer winding in the output emitter-follower stage and applying it as part of the d.c. base bias of the oscillator transistor. The a.c. signal which is to be rectified must be sufficiently large to go considerably above the knee of the rectifying diode. A large resistor can be included in series with this rectified signal to limit the derived base current. The normal d.c. bias derived from the potentiometer 'bleed chain' RV_2 sets the working conditions and hence the amplitude of the oscillator. A single limiting stage is included to produce the necessary switching square-wave for the phase discriminator. This is also fed from the output transformer primary. The inherent action of the discriminator is also that of limiting, so that the switching waveform is improved.

The three front panel controls are:

(1) A function selector switch with provision for checking battery voltage.

(2) A variable potentiometer which sets the integration zero level.

(3) A variable resistor in order to provide in-phase balance of the bridge circuit. There are no components included to give quadrature balance as this is obtained by balancing the coils themselves, by positioning the soft iron core relative to the coils, see Fig. 7. The sensitivity of the instrument can be varied or pre-set to a given value, within limits, by manipulation of controls (2) and (3). This changes the working slope of the bridge characteristics, which are non-linear.

Operation of the Instrument

In operation three separate coils are used, two contained in a probe and one in the plug on the instrument end of the probe cable. The two coils in the probe can alternate their functions, one being used as 'search' coil and the other as 'standard' coil. Fig. 8 shows the two conditions, the facility being provided so that in cases of difficult access a bolt hole may have both the forward and reverse faces searched from the same side. The third coil in the plug is free from external influences and its reflected impedance remains constant under all test conditions. It can be used as the 'standard' coil and either probe coil then used as the 'search' coil. It is normal to use the coil situated at the free end of the probe. The instrument is used as a comparator and thus a test specimen is compared against a known specimen. The double coil probe technique, employing the two probe coils, overcomes any difficulties encountered if the periphery of the test specimen bolt-hole is different to that of the known specimen. Any differences are eliminated as the two coils, in opposite arms of the bridge, are equally affected and cancel each other. It must first be ascertained, however, that there are no flaws opposite the portion of the bolt hole where the 'standard' coil is situated or erroneous results could be obtained. The single coil probe technique, using only one of the probe coils, can be used to check this condition. The conditions of the bridge remain constant ignoring end-effects of the hole, as the probe is progressed through the hole except where a flaw occurs. In this way the exact position of a flaw can be ascertained. It is normal for reaming to take

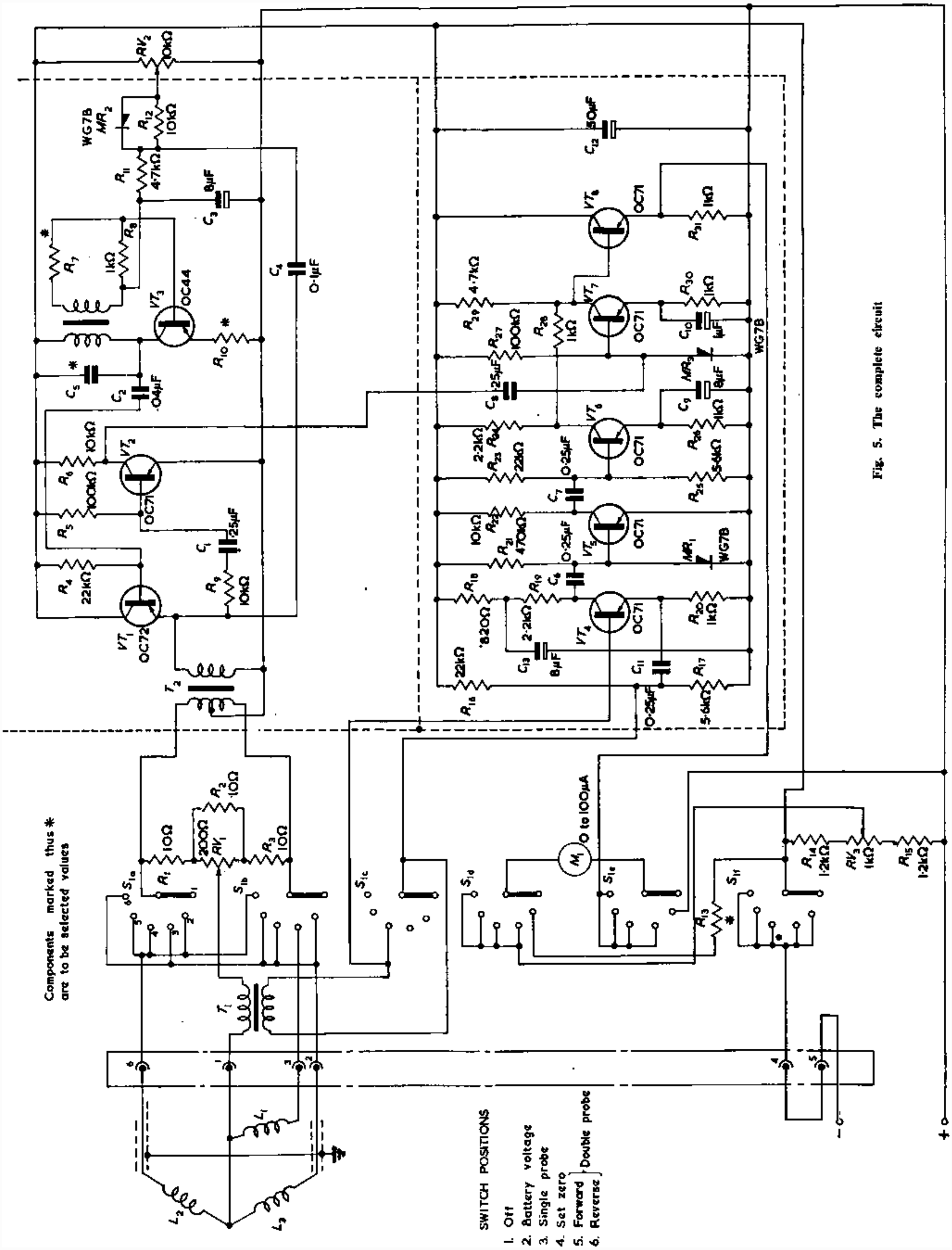

Fig. 5. The complete circuit

place prior to testing to make the bolt hole as free from minor scores as possible. The known comparator specimen must contain both good holes, i.e. free from flaws, and holes with either a known fatigue crack or a simulated fatigue crack in order to calibrate the instrument. Stringent operating instructions have been devised in order to remove, as far as possible, errors introduced by individual operators.

Further Applications

Although only one application of the instrument has been discussed in detail, there are other purposes for which it may be used. These uses are still in the development stage, but present results seem to show that the instrument is suitable for measuring paint thicknesses from 0·005in to 0·05in on aluminium alloy skins. It has also been used to test heat-treated alloys for both proof stress and ultimate tensile strength. Accuracies of the order of 8 per cent can be achieved for 24 s.w.g. materials and relatively higher accuracies as the thickness of the material increases. One practical application of the above has been the testing of fire damaged panels in an aircraft without removal and destructive testing.

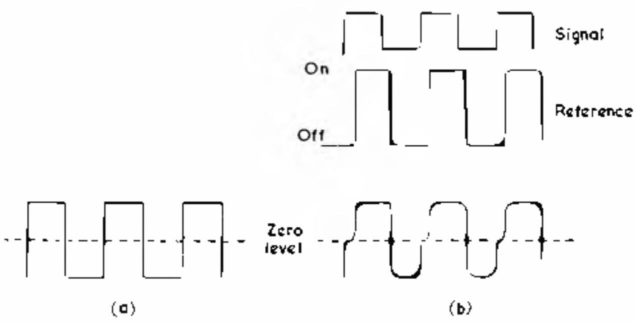

Fig. 6. Phase discriminator collector waveforms

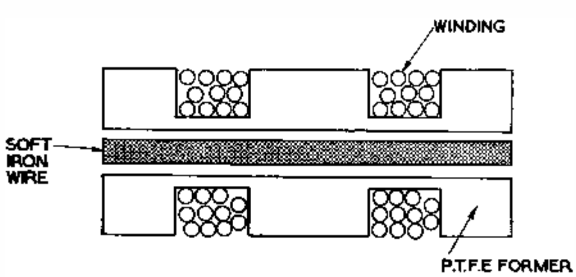

Fig. 7. Probe Layout

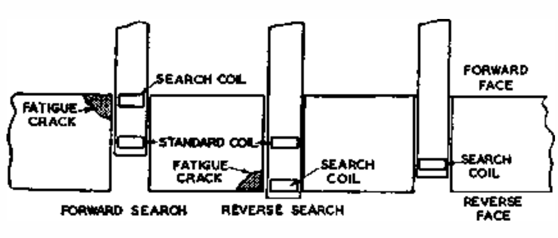

Fig. 8. Section through block containing bolt-holes

Acknowledgments

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Rigid Printed Circuits

This new method for manufacturing printed circuits relies on the Kanigen nickel phosphorous chemical deposition plating process the licence rights of which are held by Albright and Wilson (Mfg) Ltd and Societe Europiene de Revetement Chimique throughout Europe and many other parts of the world.

By its means all circuits, however complex, are produced as channels and holes in a rigid phenol-formaldehyde base, normally about $\frac{1}{8}$in thick, moulded by conventional methods. The Kanigen nickel-phosphorus metal is deposited in the channels and through the holes only using a proprietary technique of chemical deposition, after masking the remainder of the base with a chlorinated rubber paint. The Kanigen metal is deposited as a uniform, conductive layer. Plating through the holes avoids the use of 'jumpers' or other separate connectors.

Kanigen plated circuits show a better metal to base bond strength than conventional printed circuits and this factor together with the rigidity of the base eliminates the possibility of exfoliation.

Kanigen is very hard and therefore is suitable as a conductive lining for valve sockets which can be previously moulded into the circuit. The rigid base can be used as the chassis of equipment without any additional support; giving a robust, compact assembly.

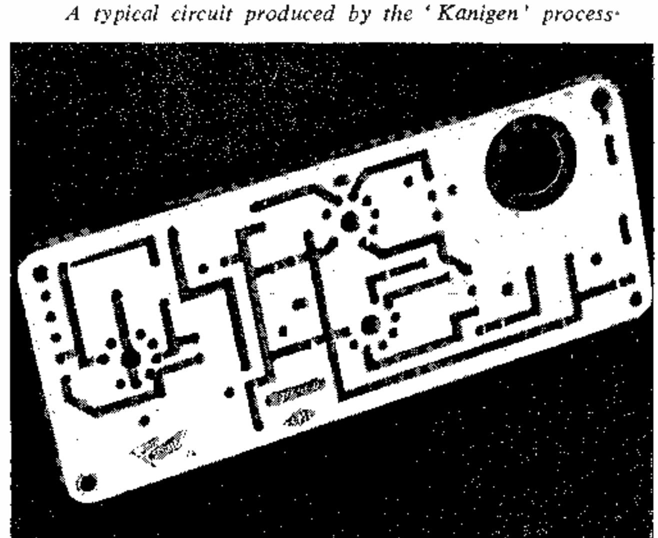
The circuit can be readily dip-soldered by using conventional techniques and shows no tendency to separate during this operation.

Silver or copper can be deposited on to the Kanigen by electro-deposition methods should this be desired.

As the corrosion resistance of Kanigen nickel phosphorus plate is greater than that of copper, particularly in tropical climates, and in contact with other metals, the risks of short-circuits caused by corrosion products is minimized. As the circuit is set below the general level of the base, short circuiting caused by dust or condensation is also less likely to occur.

It is possible to arrange a mechanical continuous flow transfer of moulded circuits throughout all stages of manufacture. The moulded bases may pass in a continuous line through the pretreatment tanks as each stage has an equal immersion time. Masking may be carried out by passing the bases on a moving belt under rollers automatically fed with a chlorinated rubber paint and thence to an infra-red drying tunnel. The masking paint may be removed after plating if necessary on a flat emery cloth belt. The masking compound is of course an added corrosion protecting and insulating material.

A typical circuit produced by the 'Kanigen' process·



A Transistor Switching Circuit for Power Regulation Applications

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

A d.c. transistor amplifier circuit for power control applications is described. Positive feedback is used to make the amplifier oscillate continuously, the output stage being switched between bottomed and cut-off conditions. The input signal modulates the mark-to-space ratio and so varies the mean current fed to an inductive amplifier load. The primary application considered for the circuit is voltage control of shunt fed dynamos or alternators, but its use in mains or battery driven regulated power supplies is briefly mentioned.

The principal advantages of a switching amplifier are (1) high efficiency and (2) low power dissipation in the series control transistor.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

THE circuits to be described were developed for use on automobile dynamos or alternators, but they can be applied to other power control problems, e.g., mains energized d.c. power supplies.

Considering the automobile application, a rotating machine driven by the engine is used to supply the electrical load when the vehicle is in motion and a means of controlling its output is required. The principle variables affecting the system are:

(1) The electrical load. This includes the storage battery which should be kept under charge at a rate depending on its state of charge.

(2) Engine speed.

(3) Temperature.

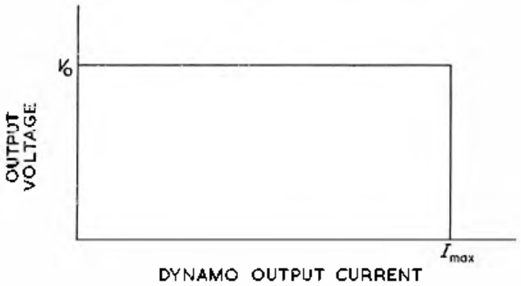


Fig. 1. Ideal regulation characteristic

The most effective control characteristic is that shown in Fig. 1 where, for a given temperature, the system voltage is maintained constant at a value V_0 somewhat greater than the battery open-circuit terminal voltage. To prevent overload of the machine, a change to constant current control is made under heavy loading conditions where the maximum rated output current I_{max} would otherwise be exceeded. Output voltage should be independent of engine speed but should fall gradually with increasing temperature to maintain a constant trickle charge current to the battery (assumed fully charged) whose internal resistance decreases with increasing temperature.

An additional control function is to disconnect the machine from the load under low engine speed conditions to prevent discharge of the battery into the armature winding. A block diagram is given in Fig. 2.

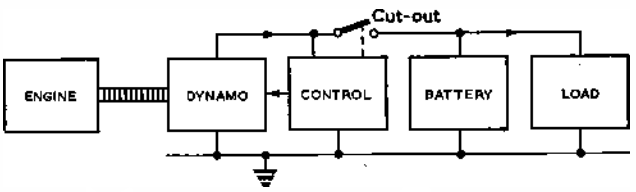


Fig. 2. Automobile electrical system

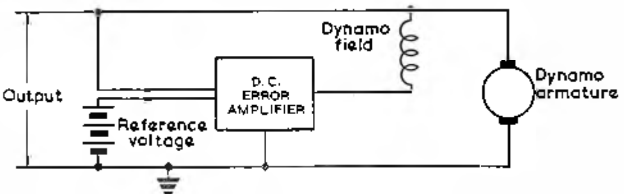


Fig. 3. Basic voltage control loop

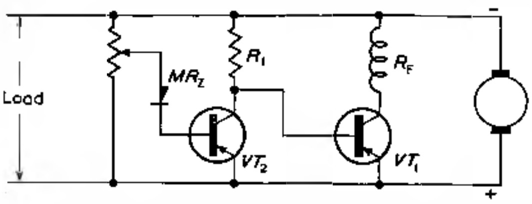


Fig. 4. Simple control circuit

In addition to its storage function, the battery greatly attenuates voltage ripple due to:

(1) Commutation (d.c. dynamo).

(2) Rectification of a.c. (alternator).

(3) Field current variations where on/off switching is employed.

* *Joseph Lucas Group Research Centre, Birmingham*

D.C. Control Amplifier

The basic control loop is shown in Fig. 3 and follows the usual pattern of comparing the output voltage with an internal reference voltage, amplifying the difference and feeding to the control winding in the required sense to decrease the error.

Fig. 4 shows a voltage control circuit using a simple d.c. error amplifier and a Zener diode for reference purposes. At low dynamo speeds, the output voltage is low so that reference diode MR_2 and transistor VT_2 are non-conducting. The output voltage is applied to R_1 and VT_1 input diode in series, the current flowing being sufficient to bottom VT_1 collector and apply maximum field excitation. When, due to an increase in dynamo speed, the output voltage approaches the desired value, MR_2 conducts a base current from transistor VT_2. An amplified current flows in VT_2 collector circuit so that the current flowing through R_1 is now shared between VT_1 base and VT_2 collector. At a certain value of VT_2 base current, VT_1 base current is reduced to a level such that VT_1 is taken out of the bottomed condition. The dynamo field current is reduced and regulation ensues.

Two values of VT_2 base current can be specified as follows:

i_{min} – The input current required to just keep VT_1 bottomed.

i_{max} – The input current required to cut off VT_1

The difference $(i_{max} - i_{min})$ is given by the maximum field excitation current divided by the current gain of the amplifier. Under the conditions of regulation, the input current i must lie between the values i_{min} and i_{max}.

Since a fraction of the output voltage is compared with the reference, the regulated output can be set to any value from $(V_z + V_t)$ upwards where:

V_z = Zener breakdown voltage of MR_z

V_t = Base-emitter voltage drop of VT_2 input diode.

Advantages of this d.c. amplifier are:

(1) Simplicity. Since the amplifier forms part of a feedback control loop, temperature stabilization components are not required.

(2) The collector-emitter voltage drop across VT_2 is always small so that a low power transistor may be used. Typically, a 10W output stage can be controlled by a 50mW input stage.

(3) At low dynamo speeds, the circuit configuration is the most effective possible for producing maximum field excitation. There are no losses (i.e. due to other transistors, diodes or resistors) in either the base-emitter or collector-emitter circuit of the output stage other than those which must be present if a transistor is to control the field winding.

(4) The circuit may be designed such that the difference $(i_{max} - i_{min}) \ll i_{min}$. This leads to a high sensitivity since i_{min} takes MR_z and VT_2 input diode into regions of low slope resistance. The change from i_{min} to i_{max} is therefore achieved by a very small change in output voltage.

In practice, the combined diode slope resistances may be too low for safety since the dynamo field time-constant prevents the control loop from responding rapidly to a sudden change in load or speed. In particular, if the dynamo load is suddenly reduced, the output voltage rises since the same field current flows instantaneously. Destruction of the Zener diode or transistor could result unless some series resistance is included. The resistance of the output potential divider can be chosen to serve this purpose.

TABLE 1
Suitable Silicon Transistor Characteristics

Base-emitter voltage (V)	0·2	0·4	0·6	0·8	1·0	1·2
Base current (mA)	—	0·1	10	50	125	250
Collector current (A)	—	—	0·1	0·7	2·5	5·0
Collector bottoming voltage (V)	—	—	—	—	0·4	0·75

No difficulty has been encountered in practice with the direct connexion between VT_2 collector and VT_1 base since the bottoming voltage of a low power germanium transistor is small enough to control the power germanium transistor over the desired range of output current. Should any doubt exist, a silicon diode can be added in series with the base of VT_1. Providing that VT_2 bottoming voltage is less than say 0·7V, VT_1 can then be completely cut off by VT_2. The addition of a resistance between the base and emitter terminals of the power transistor will prevent a complete open-circuit base condition from occuring. If a silicon power transistor is used for VT_1 no additional diode is required. However, if the transistor is to supply the maximum field current efficiently it must have a low collector saturation resistance. The characteristics of a typical pre-production silicon transistor (npn type) developed for machine control are given in Table 1.

For this 'continuous' control circuit the maximum power dissipation P_{max} in the output stage is given by:

$P_{max} = V_o^2/4R_F$

V_o = Regulated output voltage

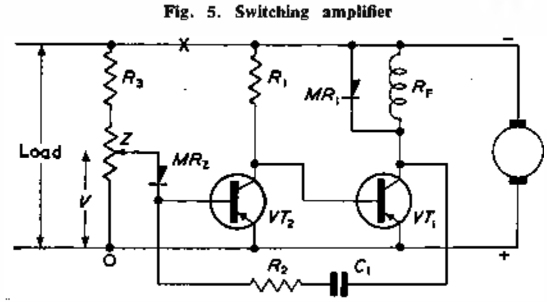
R_F = Field resistance

Typically with V_o=15V (nominal 12V system) and $R_F = 6\Omega$, P_{max} = 9·4W.

Switching Control

A much reduced dissipation in the output transistor can be achieved by causing it to switch at a low audio frequency between the bottomed and cut-off conditions. The mean field current is varied by alteration of the mark-to-

Fig. 5. Switching amplifier



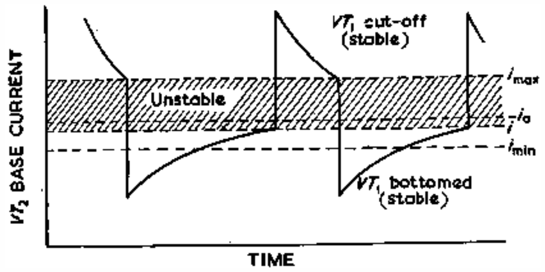


Fig. 6. VT_2 base current variations (Fig. 5 circuit)

space ratio, the dynamo field time-constant smoothing out field current variations.

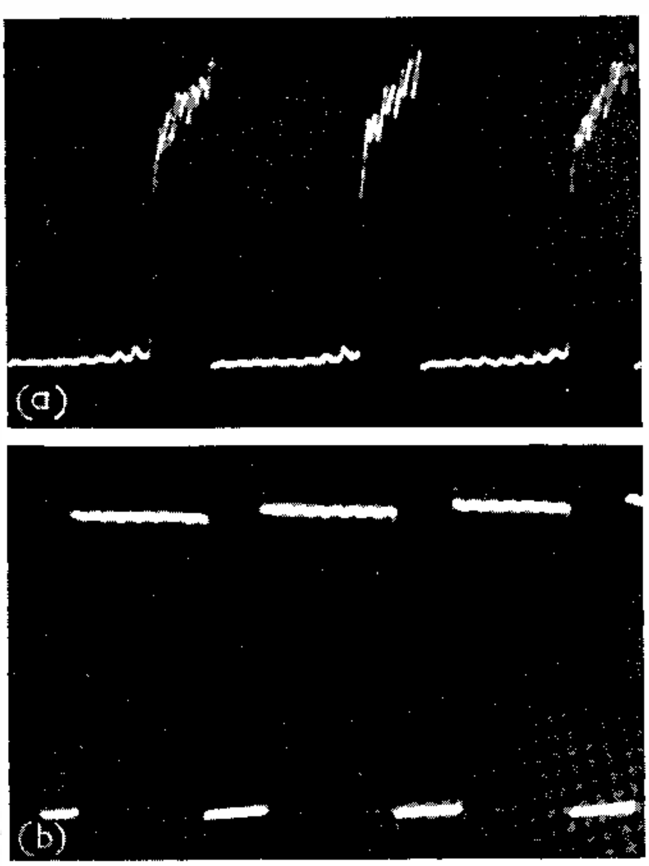
The circuit of Fig. 4 can be converted to this form of control by the addition of a.c. positive feedback as shown in Fig. 5. Given suitable component values, the amplifier is unstable within the regulation range and oscillates freely. A diode is required in parallel with the field winding to maintain the flow of field current when VT_1 cuts off.

Suppose that under given operating conditions the field current required is produced with an amplifier input current of value i (see Fig. 6). With VT_1 in the cut-off condition, a charging current for capacitor C_1 holds the input current greater than i_{max} which keeps VT_1 cut off. However, the charging current decays exponentially towards a mean value i_a, which lies somewhere between i and i_{max}, until it reaches the value i_{max}. VT_1 now commences to conduct and field current originally flowing entirely through diode MR_1 divides between MR_1 and transistor VT_1. The fall in voltage across diode MR_1 is coupled back to the input terminal via C_1 and R_2 and a cumulative action occurs resulting in bottoming of VT_1.

Fig. 7. Waveforms for Fig. 5 circuit

(a) VT_1 collector current against time

(b) VT_1 collector to emitter voltage against time



The input current now rises exponentially towards the mean value. When the input current reaches the value i, transistor VT_1 can no longer support the field current flowing. Since VT_1 load is inductive, any reduction in current causes a large change in voltage at VT_1 collector and a rapid switch off occurs.

The machine output voltage still controls the average input current to VT_2. Any change in dynamo speed or load leads to a change in the average current i_a and a change of mark-to-space ratio to suit. No switching can occur under conditions where the dynamo cannot support the voltage V_o across the load since the input current will be below the value i_{min} (a stable region). Thus maximum excitation is still applied to the dynamo until the region of regulation is reached.

It will be seen from Fig. 6 that for satisfactory switching of the output stage over the full regulation range, the change in VT_2 base current due to feedback must exceed the value $(i_{max}-i_{min})$. Not all the current flowing through R_2 flows into VT_2 base, of course, since the slope impedance of the Zener diode plus the output resistance of the potential divider network are in parallel with the slope impedance of VT_2 input diode.

Fig. 7 shows oscillograms of VT_1 collector current and VT_1 collector to emitter voltage against time for the circuit of Fig. 5, the frequency of oscillations being 60c/s. A storage battery and a resistive load were present across the dynamo armature. Frequency components due to commutation as well as the switching process are present on the output voltage and dynamo field current.

The choice of oscillation frequency is limited at low frequencies by the ripple produced at the output terminals while the upper frequency limit is set by the mean power dissipation in VT_1 due to the switching transients. The frequency is determined principally by components C_1 and R_2 and the amplifier gain.

Transistor Dissipation

The maximum mean power dissipation in the output transistor now occurs at the onset of regulation when the transistor carries the maximum field current in the bottomed condition. For $V_o=15$V, $R_F=6\Omega$, and assuming a transistor saturation resistance of 0·15Ω, the maximum excitation current I is given by: $I=15/(6+0{\cdot}15)$ and the maximum power P_{max} by $P_{max}=I^2R=(15/6{\cdot}15)^2\times0{\cdot}15=$ 0·9W.

Dissipation due to switching transients can be reduced as follows:

(a) *Switch on*

The addition of a resistor in series with MR_1 gives a larger change in feedback voltage for a given reduction in diode current.

(b) *Switch off*

A parallel CR network between VT_1 base and the junction of R_1 and VT_2 collector enables VT_1 base to be driven positive with respect to its emitter when VT_2 bottoms. This gives a more forcible clearing of current carriers in the base region.

The reduced power dissipation due to switching allows one or more of the following advantages to be gained.

(1) Use of a lower power transistor for VT_1.

(2) Use of a smaller heat sink.

(3) Operation at higher ambient temperatures.

An additional advantage is a higher machine efficiency. Under non-switched conditions the field winding bleeds the required excitation current continuously from the armature output current. Under switched operation this same value of current is taken during switch on only.

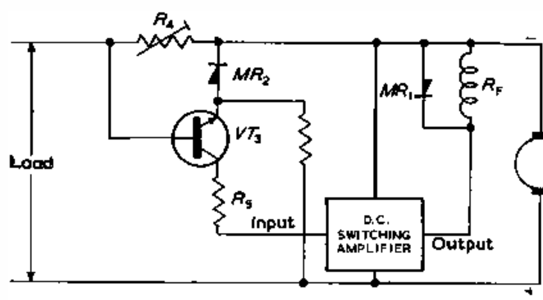


Fig. 8. Current limiting circuit

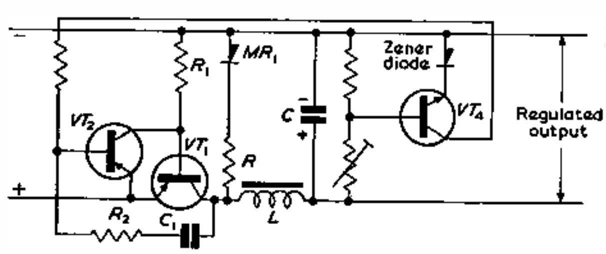


Fig. 9. Power supply regulation circuit

Temperature Compensation of Voltage

A thermistor is easily incorporated in the resistive potential divider network (in parallel with R_3, Fig. 5) to give an output voltage dependent on ambient temperature. Increasing temperature gives a reduction in thermistor resistance (R_T). Diode MR_2 regulates the potential difference V between points o and z which, for a given potentiometer setting, gives a constant current through the network (ignoring reference diode current). The output voltage is therefore proportional to the total network resistance and follows variations in the parallel combination of R_T and R_3. By appropriate choice of R_T and R_3 a number of voltage/temperature characteristics are obtainable. For example, a linear fall of 1·5V over the temperature range 0°C to 60°C is achieveable to an accuracy of $\pm$0·1V.

Current Control

A current limiting circuit is shown in Fig. 8. The load current flows through resistor R_4 and the potential difference across it is compared with that across the reference diode MR_2, the difference being applied between the base and emitter of npn transistor VT_3. At an output current such that the voltage across R_4 exceeds that across MR_2 plus the forward voltage drop of VT_3 base-emitter diode, an input current flows in VT_3. VT_3 collector current is applied, via current limiting resistor R_5, as an input current to the switching amplifier previously described. MR_2 can be a silicon junction diode and the forward characteristic used (as shown) or could be a Zener diode in applications where a few volts drop in resistor R_4 is permissible.

The current and voltage regulators can be combined, using the same amplifier, to give a control characteristic of the type shown in Fig. 1.

When the rotating machine is an alternator no current control circuit is required since the machine can be designed to give self-limitation at the required current.

Cut-Out Diode

A series diode positioned between the dynamo output and the battery prevents the battery from discharging into the dynamo armature when the engine is stationary or at tick-over speeds. It is best positioned within the voltage regulator control loop (point x in Fig. 5) where its voltage/current characteristic has negligible effect on overall performance.

Since an automobile alternator includes diodes for rectification purposes, no additional cut-out diode is required.

Regulated Power Supplies

High efficiency regulated power supplies result from the use of a switching amplifier in place of the usual d.c. control amplifier. One such circuit is shown in Fig. 9. Transistor VT_1 periodically connects the unregulated supply to the load via an inductor L. During the periods when VT_1 is non-conducting, the inductor maintains the current supply via diode MR_1 and resistor R (speed up resistor for switch-on). Transistor VT_4 is the usual voltage error detector and feeds the switching section, stages VT_1 and VT_2. No continuous power dissipation is lost in the series control transistor. During the on period, the unregulated supply feeds the load and stores energy in the inductor. During the off period no current is taken from the unregulated supply. Thus the average current drawn from the unregulated supply is less than the load current by some factor depending on the mark-to-space ratio. This leads to a high power conversion efficiency even where the regulated voltage V_r required is only a fraction of the unregulated supply V_u. For the more usual continuous control circuit, the maximum possible efficiency is limited to V_r/V_u.

The choice of value for the inductor L depends on whether rapid changes in output current are desired. High values of inductance, giving minimum output ripple, may be used when the load current varies relatively slowly. Lower values of inductance should be used to allow fast changes in output current with minimum transient effects on output voltage.

The switching frequency should be as high as possible without allowing transient dissipations to contribute more than a fraction towards the mean dissipation in the output stage VT_1.

In common with the usual continuous control curcuit, the switching circuit has an appreciable filtering action on any ripple voltage at the output of the unregulated supply.

Acknowledgment

The author wishes to thank the Directors of Joseph Lucas Ltd, for permission to publish this article.

Phase Sensitive Demodulation with Interpolation

By J. S. Johnston, B.Sc., A.R.C.S., A.Inst.P.

This article describes a method of obtaining a very much better approximation to the envelope waveform of a modulated, low frequency carrier, than is possible by conventional phase sensitive demodulation. The method depends on some previous knowledge of the nominal shape of the waveform, and particular attention is given to the important sinusoidal case.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

PHASE sensitive demodulators find one of their main applications in recovering the modulation envelope, from the signal returned from an a.c. resolver of some kind.

A typical signal of this type is shown in Fig. 1 in which the modulation envelope is indicated by the heavy bounding line.

There are a variety of techniques for performing the demodulation[1], but all of them apply to the case that the modulation frequency (rotation frequency of the resolver) is very small compared with that of the carrier. While in most systems this is not difficult to arrange, there are others in which a signal may be required from a fast moving resolver and yet some other factor (for example

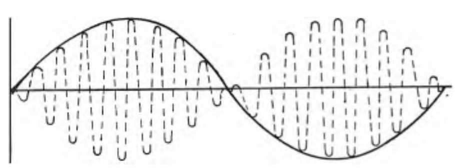


Fig. 1. Typical output from resolver, showing ideal modulation envelope

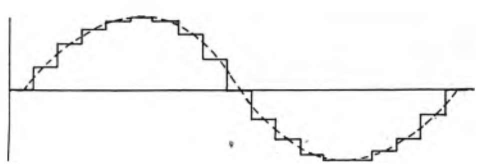


Fig. 2. Response of demodulator to waveform of Fig. 1, with no interpolation

the need to drive the resolver at the end of a long length of cable) prohibits the use of a high carrier frequency.

The interpolation method described here was devised to overcome this difficulty in the case of high speeds of revolution of a remote radar aerial.

Demodulation by Pulse Sampling[2]

Before considering the interpolation process, it is necessary to describe the particular type of demodulator used in this work. This is of the form known as a 'pulse-sampling' demodulator, and operates by storing for one period of the carrier wave, the instantaneous potential of the waveform at the last peak of the carrier; the resulting output is indicated by the solid line in Fig. 2. It will be noted that the smooth envelope of Fig. 1 has been quantized by this process into discrete steps. The relation between the two frequencies controls the coarseness of the quantization, one result of which, as can be seen from the dotted curve, is that on the average, the output has a phase lag with respect to the true envelope which corresponds to half a carrier period.

Fig. 3 is a simplified diagram of the demodulator circuit which formed the basis of this method. Its operation is described below and the reason for its apparent complexity will become clear later.

Consider the point Y to be held at a negative potential

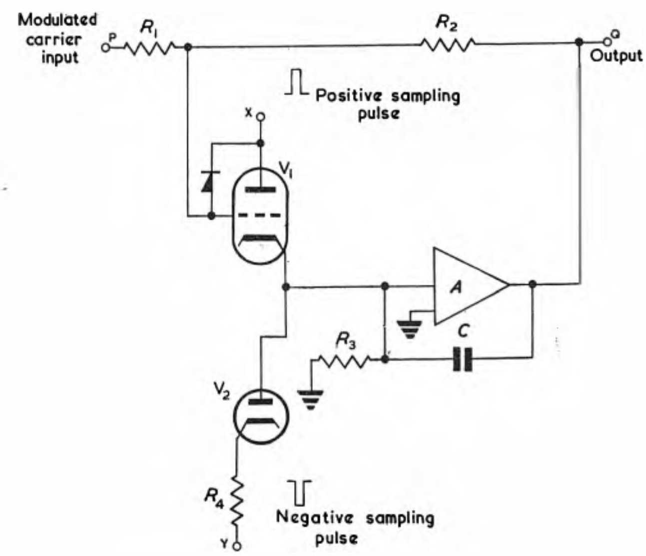


Fig. 3. Simplified diagram of pulse-sampling demodulator

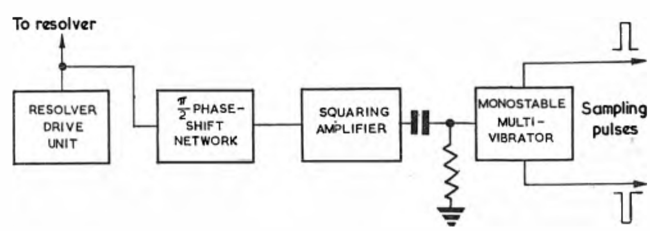


Fig. 4. Generation of sampling pulse

and the point X at a positive potential. The triode V_1 acts as a cathode-follower with the diode V_2 and resistor R_4 as cathode load. This arrangement presents a low impedance to the input of the high gain amplifier A which is shunted by a capacitor C. The entire circuit, with input at P and output at Q, now behaves as a sign changing amplifier with gain R_2/R_1 (assuming the gain of A to be very large); the low output impedance of V_1 ensures that C can charge and discharge fast enough to follow the input waveform.

If now, the potentials on X and Y are reversed, both V_1 and V_2 will be cut off, presenting a very high impedance at the cathode of V_1. The components R_3 and C, together with the amplifier, form an integrating circuit.

Since R_3 is earthed, however, there is no flow of current into C and hence no change in the output voltage of the amplifier; the output therefore must remain at the voltage

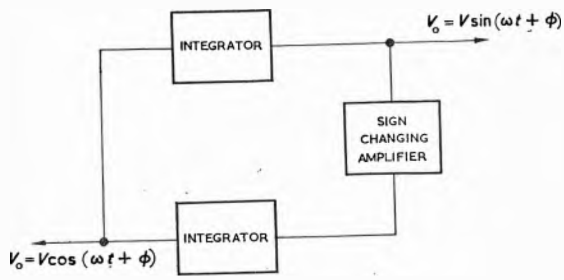


Fig. 5. V.L.F. oscillator loop

which existed there just before the potentials on V_1 and V_2 were reversed.

In operation as a demodulator, the circuit is switched to the first condition only during a brief sampling interval at the peak of the carrier waveform; the sampled voltage (reversed in sign and amplified in the ratio R_2/R_1) is then stored on C until the next sample is taken.

Generation of Sampling Pulses

One method of generating the required sampling pulses can be seen from Fig. 4. The driving waveform for the resolver is phase shifted by $\pi/2$, this signal passes into a squaring amplifier whose output is used to trigger the pulse generating flip-flop. The phase shift ensures that the sampling pulse coincides with the peak of the carrier.

Smoothing of Output by Interpolation[3]

Passive filtering of the step waveform of Fig. 2 will not give useful smoothing unless the time-constant of the filter is comparable with a period of the modulation envelope; if it is of this magnitude the additional phase shift and attenuation produced will usually prove intolerable.

However the general form that the modulation envelope should take is very often known in advance, this knowledge if available, allows precise interpolation of the missing parts of the envelope to be made. In the case quoted above it might be known for example, that the aerial was designed to rotate at 60rev/min ± 10 per cent. Under these circumstances there appear at the two outputs from the resolver, modulation envelopes which are sinusoids of this frequency and 90° out of phase with the other.

The problems of interpolating between samples then reduces to constructing an oscillator of the nominal frequency, having two outputs in phase quadrature and correcting these outputs in phase and amplitude at each sampling pulse.

L.F. Oscillator Loop

Fig. 5 shows the familiar computing loop for the simple harmonic equation.

$$V = -\frac{\omega^2 d^2 V_o}{dt^2} \qquad \text{(1)}$$

At the two outputs of this circuit, the voltages $V \sin(\omega t + \phi)$ and $V \cos(\omega t + \phi)$ appear, in which ω is set by the time-constants of the integrators while V and ϕ are decided by the initial conditions.

Fig. 6 shows this loop combined with the pulse-sampling demodulators. The natural frequency of oscillation is set to be the same as that of the nominal modulation envelope, and each sampling pulse resets the output to the instantaneous phase and amplitude of the envelope.

The waveform produced is shown in Fig. 7. Obviously, if the oscillator frequency were just correct, even the small voltage steps shown here would not be present.

Improvement in Reproduction of Envelope

It will readily be seen that in the case of simple 'step and hold' demodulation by a conventional phase sensitive demodulator, the approximation to the envelope is most in error just before a fresh sampling pulse resets the output. This angular error is

$$\Delta\,\theta_1 = \frac{360\ f_a}{f_o}\ \text{degrees} \qquad \text{(2)}$$

where f_a is the rotation rate (rev/sec)

and f_o is the carrier frequency in (c/s).

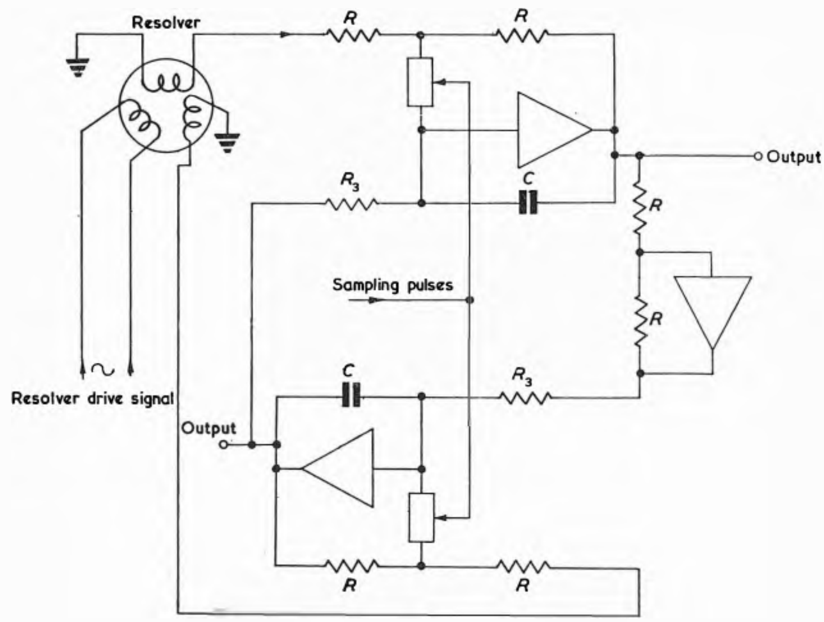


Fig. 6. Complete demodulator

When the above interpolation scheme is in use, and the integrator time-constants are such as to set an oscillator frequency of f_1 cycles per second this error reduces to

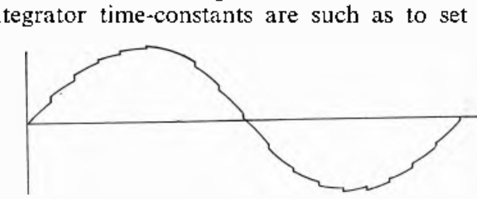


Fig. 7. Response of demodulator to waveform of Fig. 1, with interpolation in use

$$\Delta\,\theta_2 = \frac{360\ (f_a - f_1)}{f_o}\ \text{degrees} \qquad \text{(3)}$$

To take a particular example, if the shaft were to rotate at a nominal 60rev/min with a 400c/s drive to the resolver then $\Delta\theta_1$ would be 0·9°. Using an oscillator frequency of 60c/min this error is reduced to the order of 0·09° or less over an actual variation in shaft speed from 50 to 70 rev/min when interpolation is used.

Other Modulation Envelopes

The example above has been concerned with the particular case that the modulation envelope is sinusoidal. It is of course possible to use a similar system for any modulation envelope whose nominal characteristics are known, and for which an analogue computing loop can be constructed. It is obviously essential that each sample should contain enough information to enable all the relevant boundary conditions to be established for the next period.

This method of demodulation has been successfully used to extract aerial bearing information from resolvers at remote sites without the need for local servo repeaters.

Other Fields of Application

The method has an extension to analogue computing generally in that, by the use of these fast electronic switching techniques, known spot values can be fed into the computer during the running of a problem in order to minimize the effect of long term drift errors.

Conversely the same method can be used to provide analogue interpolation between discrete values obtained from a digital computer or process controller.

Acknowledgments

The author in indebted to Mr. R. Matthews for the original suggestion that led to this work and to Mr. **R. F.** Hansford of Decca Radar Ltd for permission to publish this article.

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A Simple Supply-Compensated Constant Current Source

By P. Pargas*, M.I.E.E.

A simple circuit is described to provide a current of very high stability for the measuring slidewire of a self-balancing potentiometer. The overall long-term stability obtained over a period of 1 000 *hours is better than* 0·15 *per cent, in spite of mains fluctuations and temperature variations.*

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

ALTHOUGH a large number of articles have appeared on various aspects of stabilizing circuits and their applications, only a few have dealt with simple stabilizing circuits using only one or two non-linear elements. Unfortunately these give little detailed care to the problem of compensation against mains fluctuations.

The requirements for a particular slidewire supply circuit are described below, but the principles established are quite general and can be applied to widely varying requirements.

The self-balancing potentiometer under discussion has an overall accuracy of ±0·5 per cent, but it is desirable that the stability of the slidewire supply should be ±0·15 per cent or better. The load on this supply is 6mA at about 1V.

Simple Stabilizing Circuits

Simple stabilizing circuits using only one cold-cathode stabilizing tube, or two tubes in cascade, are well known, but are included here to show their limitations. These arise from the fact that all stabilizing tubes have a finite incremental resistance. Thus the output of such circuits always varies with changes in the mains voltage.

Consider now a simple stabilizing circuit as shown in Fig. 1.

Here R_o represents the slidewire and the appropriate series resistances. Replacing the stabilizing tube by an equivalent constant voltage generator V with an internal impedance r_a (corresponding to the incremental resistance of the tube), it can be shown that the load current is:

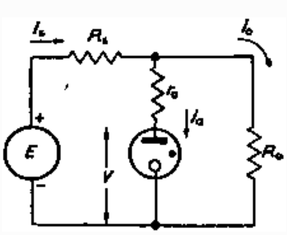
$$i_o = \frac{E \,.\, r_a + V \,.\, R_s}{r_a \,.\, R_s + r_a \,.\, R_o + R_s \,.\, R_o} \quad \ldots\ldots\ldots\ldots \; (1)$$

and the stabilized voltage across the load is:

$$v_o = \frac{R_o\,(E \,.\, r_a + V \,.\, R_s)}{r_a \,.\, R_s + r_a \,.\, R_o + R_s \,.\, R_o} \quad \ldots\ldots\ldots\ldots \; (2)$$

Now δv_o, the change in output voltage corresponding to

Fig. 1. Series regulated uncompensated stabilizing circuit



a change $\delta E = \beta \,.\, E$ in the mains voltage, can be shown to be:

$$\delta v_o = \frac{\beta \,.\, E \,.\, r_a \,.\, R_o(i_a + i_o)}{(E - V)(r_a + R_o) + r_a \,.\, R_o(i_a + i_o)} \quad \ldots\ldots\ldots \; (3)$$

where β is the fractional change in the supply voltage. When equation (3) is represented graphically (see Fig. 2), it is found that δv_o shows an abrupt change at $E - V$, where δv_o changes over into the negative, which does not represent a real solution. For the real part, the smallest value of δv_o occurs at $E = \infty$, when:

$$\lim_{E \to \infty} \delta v_o = \beta \,.\, r_a \,.\, i_a \quad \ldots\ldots\ldots\ldots\ldots\ldots \; (4)$$

* *LogEtronic Inc., U.S.A., formerly Sunvic Controls Ltd*

This gives us the limiting values for stabilization for any one set of circuit components.

From equation (3) dividing by R_o:

$$\delta v_o = \frac{\beta . E . r_a (i_a + i_o)}{(E - V)\, r_a / R_o + (E - V) + r_a(i_a + i_o)}$$

But as $E - V \ll r_a (i_a + i_o)$ and also $r_a \ll R_o$, one can approximate and write:

$$\delta v_o = \frac{\beta . E}{E - V} (i_a + i_o)\, r_a \quad \ldots\ldots\ldots\ldots ((5)$$

i.e. the eror in the load voltage is directly proportional to the incremental resistance of the tube and also to the total current $i_a + i_o$.

To illustrate the above, Fig. 3 is given for a supply voltage of 300V and a change in it of 10 per cent for an 85A2 reference tube.

As the stabilizing element it was decided to use a Mullard 85A2 reference tube which seemed to be the most suitable for the particular requirements. (For example for an average tube the change in output current would be 0·78 per cent.)

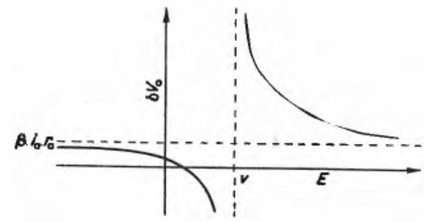


Fig. 2. Change in δv with supply voltage changes

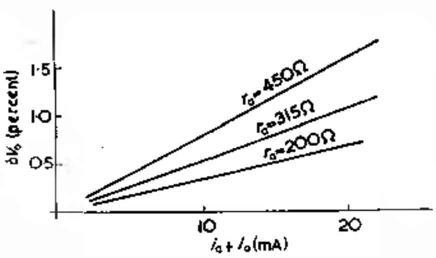


Fig. 3. Effect of internal resistance of tube on δv

To improve the stability, two tubes can be connected in cascade, the first tube normally having a maintaining voltage of 150V. For the best such tube available the long-term stability is only $\pm 1{\cdot}0$ per cent per 1 000 hours; this alone would cause a load current change of $\pm 0{\cdot}13$ per cent (see equation (3)) for a load current of 6mA. The overall accuracy then obtainable for a 300V ± 10 per cent supply is about $\pm 0{\cdot}36$ per cent. Difficulties arise, though, in ensuring the ignition of both tubes. This can only be done satisfactorily with a reduced load current of 3mA or less. Even then the stability is improved only to $\pm 0{\cdot}28$ per cent.

The above shows clearly that simple stabilizing circuits are seriously limited in their stability against changes in the supply voltage.

Supply Compensated Circuit

A circuit giving a much higher degree of stabilization has been developed and is shown in Fig. 4.

Here the stabilizing tube is represented by V and r_a. For simplicity in the calculations it is assumed that the supply is of zero impedance, this is mainly true for a practical circuit, as the source impedance is very much smaller than the impedance of other components in the circuit.

It can be shown that for this circuit:

$$i_o = \frac{(R_s + r_a) . (R_m + R_n)}{R_s . r_a(R_n + R_m) + R_n . R_m(R_s + r_a) + R_o(R_s + r_a)(R_n + R_m)} \left[V\left(1 - \frac{r_a}{r_a + R_s}\right) + E\left(\frac{r_a}{r_a + R_s} - \frac{R_n}{R_n + R_m}\right) \right]$$

and the rate of change of load current with respect to changes in the supply voltage E is given by:

$$di_o/dE = \frac{(R_s + r_a) . (R_m + R_n)}{R_s . r_a(R_n + R_m) + R_n . R_m(R_s + r_a) + R_o(R_s + r_a)(R_n + R_m)} \left[\left(\frac{r_a}{r_a + R_s} - \frac{R_n}{R_n + R_m} \right) \right]$$

It can be seen that the load current is independent of the supply voltage, i.e. that $di_o/dE = 0$, only when:

$$r_a / R_s = R_n / R_m$$

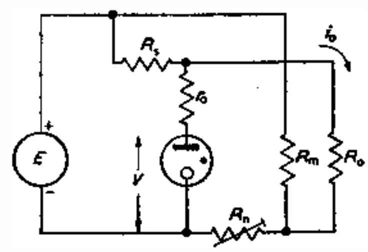


Fig. 4. Supply compensated stabilizing circuit

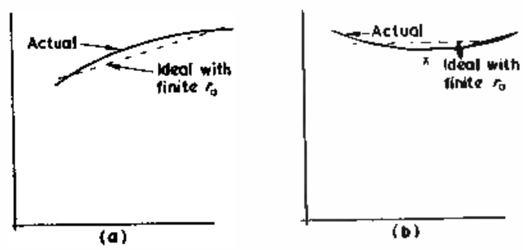


Fig. 5. Regulation of (a) uncompensated, (b) compensated circuit

If a circuit could be designed to satisfy this equation, an ideal stabilizing circuit would result.

Consideration will now be given as to how this problem can be approached in practice for the best results.

Choice of Operating Current

(a) The incremental resistance of the 85A2 over its working range is generally not constant. It is at its highest at low currents and decreases towards higher currents, and tends more towards a constant value as the current increases. The output of the circuit will therefore vary as shown in Fig. 5.

Fig. 5(a) shows the characteristic of the simple stabilizing circuit, while Fig. 5(b) shows the compensated circuit. (At point x on actual curve $r_a / R_s = R_n / R_m$). Both stress the importance of using the tube over its most linear range.

(b) The characteristic of the 85A2 shows steps of up to 50mV, most of which occur below 5·5mA. The frequency of occurrence and their amplitude depend on the physical construction of the tube and vary from tube to tube. The reason for these discontinuities is the spreading of the cathode glow in steps. This phenomenon restricts the useful range from 5·5 to 10mA.

(c) The stability of the 85A2 decreases with increase in current, but this decrease is not very large over the working range of the tube.

The factors considered in (a) and (b) outweigh the importance of (c). The operating current of the tube has therefore been fixed at 7mA.

Effects of Temperature

The temperature coefficient of the 85A2 is -4mV/°C, this corresponding to 0·005 per cent/°C. For a normal ambient change of ± 15°C this would represent a change of $\pm 0{\cdot}075$ per cent in load current. For more exacting circuit requirements, however, a temperature sensitive element with positive temperature coefficient can be made part of R_m.

Hysteresis

Any hysteresis effect of voltage with respect to current is exhibited for the 85A2 only at very low currents and also at higher frequencies. Since d.c. only is dealt with, and only over the higher current range of the tube, hysteresis effects can be neglected.

Long-term Stability

For the 85A2 reference tube the manufacturers quote a long-term stability of $\pm 0{\cdot}1$ per cent for 1 000 hours (after the first 300 hours), this figure, however, only applies to a circuit where the current through the tube is kept constant at the preferred value of 6mA. No information from the manufacturers for using the tube at a varying current is so far available. Experimental results taken for a number of tubes show, however, that over the current range used of $7{\cdot}2 \pm 1{\cdot}2$mA, the stability can be taken to be the same as that for the preferred standing current.

Inherently the compensated circuit has a somewhat worse long-term stability than the simple stabilizing circuit. Consider a change in load current due to changes in V and r_a.

$$\delta i_a = (\delta i_a/\delta V)\delta V + (\delta i_o/\delta r_a)\delta r_a$$

Taking $k = R_s/r_a = R_m/R_n$, where $k \gg 1$, the error in the load current can be shown to be:

$$\delta i_a \simeq \frac{1}{r_a + R_n + R_o}\,\delta V + \frac{E(r_a+R_n+R_o) - V(kr_a+R_n+R_o)}{kr_a(r_a+R_n+R_o)^2}\,\delta r_a$$

When evaluating the above for the compensated circuit for a 0·1 per cent change both in V and r_a, one obtains:

$$\delta i_o/i_o = 0{\cdot}122 \text{ (per cent)}$$

The increase is only 0·022 per cent, higher than that of a simple stabilizing circuit, but the long-term stability is not seriously impaired.

Experimental Circuit

Taking into account the above, a circuit was designed incorporating component values as follows:

$R_m = 33\text{k}\Omega$ $\qquad R_s = 15\text{k}\Omega$

$R_n = 1\text{k}\Omega$ variable $\qquad R_o \simeq 14\text{k}\Omega$

The output was measured across part of R_o at a level of 1V.

Measurements were taken for five tubes. It was attempted to cover the complete range of incremental resistances from 200Ω to 450Ω in the choice of these tubes. For each tube the compensating resistance R_n was adjusted individually for the best compensation. The changes with ± 10 per cent change of mains voltage for each tube are given in Table 1.

TABLE 1

	INCREMENTAL RESISTANCE r_a (Ω)				
	233	278	294	343	467
δi_o (%)	$\pm 0{\cdot}026$	$\pm 0{\cdot}047$	$\pm 0{\cdot}012$	$\pm 0{\cdot}037$	$\pm 0{\cdot}033$

The quality of compensation apparently does not depend on the actual value of the incremental resistance of each tube, but on the degree of non-linearity of the tube over the working range. This is the ultimate limitation in obtaining an ideally compensated circuit.

To test the long-term stability, six circuits were built up. These were all adjusted to give the best compensation using brand new tubes. Then the outputs were measured (after a 300-hour period) every 24 hours for 1 000 hours. The maximum percentage deviation from the reading at the start of the 1 000-hour period is given in Table 2.

TABLE 2

	V_{01}	V_{02}	V_{03}	V_{04}	V_{05}	V_{06}
$+\delta V$ (%)	0·07	0·14	0·07	0·02	0·08	0·11
$-\delta V$ (%)	0·11	0·12	0·04	0·10	0·12	0·03
mean	$\pm 0{\cdot}09$	$\pm 0{\cdot}13$	$\pm 0{\cdot}06$	$\pm 0{\cdot}06$	$\pm 0{\cdot}10$	$\pm 0{\cdot}07$

When incorporating the circuit into a self-balancing potentiometer, the circuit did not operate satisfactorily due to the presence of capacitive currents generated in the transformer of the supply circuit. The capacitive currents developed an a.c. potential across the input of the potentiometer. This appeared in the amplifier as unwanted noise and saturated the amplifier. To improve operating conditions, the secondary of the transformer was surrounded by a double shield. The outer shield was grounded, the inner shield was connected to the centre-tap of a balancing network across both lines (a full-wave rectifier is used). This improved the performance considerably. For even better performance, however, a C-core transformer was used with primary and secondary on separate limbs, the latter split into two separate symmetrical parts. All windings were individually shielded. As an additional precaution, the leads to the rectifiers were kept very short. The insulation between secondary and core was well above $3 \times 10^{10}\Omega$.

Conclusions

The circuit satisfies the requirements, and for the average tube is better than the requirements. It is virtually independent of surges in the supply and has a negligible time-constant, it being that of the tube.

It can be used as a relatively high (up to 75V) reference of considerable accuracy and stability, especially suitable for shorter experiments, as then the long-term stability of the tube is better (0·01 per cent per 8 hours) and consequently the circuit will be better than 5 parts in 10 000.

To maintain the required stabilization over longer periods, it is not necessary to make any adjustments.

If higher load currents are required, stabilizing tubes with higher current ratings have to be used. In this case the long-term stability of the tubes used will be the limitation to the stability of output, mains fluctuations will be negligible compared to them.

Acknowledgment

The author expresses his thanks to Mr. N. R. Davis of Sunvic Controls Ltd for his permission to publish this article and to Mr. K. Gimson of Mullard Ltd for making available information on stabilizing and reference tubes.

REFERENCE

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The Correction of Errors in Potentiometer Function Generators

By J. Merchant*, B.Sc.

A method is described of correcting the various errors that can occur in potentiometer function generators, due to arm loading and other practical limitations.

The errors caused by approximating to a continuous function with line segments are not considered. In the method, the potentiometer is loaded with resistors calculated in the most simple way, to give the required line segment function. The resulting function is measured, and its deviations from the design objective used to calculate a set of correcting conductances for the potentiometer segments. The potentiometer, modified with these conductances will now give a function much nearer the design objective. A second application of the method may be made to improve still further the function accuracy, but in many cases one correction process will be sufficient.

(Voir page 527 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 532)

A POTENTIOMETER that has a number of taps may be used to generate arbitrary functions of the angular position of the shaft. Loading resistors connected across the segments of the potentiometer can be chosen to produce the necessary voltage gradients (Figs. 1 and 2).

This method of function generation is often used in analogue computers. A typical circuit is shown in Fig. 3.

If the potentiometer arm is connected to a very high impedance it may be assumed that it will draw no current

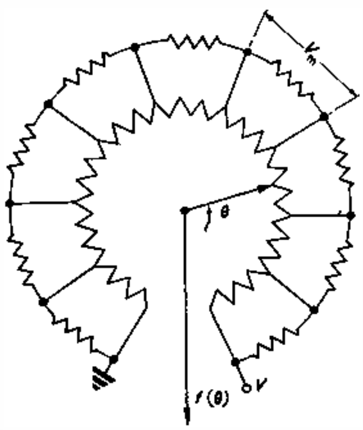

Fig. 1. Potentiometer loaded with resistors to give appropriate segment voltages

from the potentiometer. In this case the calculation of the values of the loading resistors is quite straightforward. However, it is difficult to calculate exact values of the resistors when the arm loading is not negligible.

In general the errors in a function constructed with a loaded potentiometer are due to:

(1) Loading of the potentiometer arm.

(2) Deviations of the value of the segment loading resistors from the design value.

(3) Deviations of the position of the potentiometer tap points from nominal.

(4) Deviations of the impedance of the unloaded segments from nominal.

In principle each of these sources of error could be separately eliminated by appropriate calculations and measurements. However, this would be a rather tedious method.

In this article a technique is described of correcting for all errors in one step.

Method

The segment loading resistors r_m are calculated assuming no loading of the potentiometer arm. The potentiometer is loaded with these resistors and the function measured. Let V_m' be the required potential difference across the m^{th} segment and V_m the value actually obtained. A set of extra

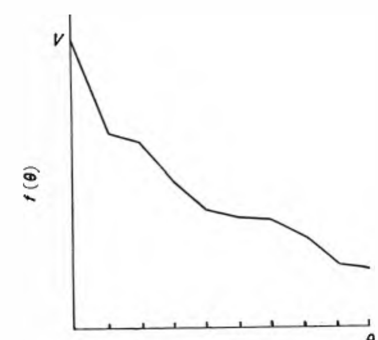

Fig. 2. Typical line segment function generated by loaded potentiometer

shunt resistors S_m is now connected across the segments where:

$$S_m = \frac{R_o V_m'}{V_o(V_n'/V_n - V_m'/V_m)}$$

and R_o is the total impedance of the segment with the largest voltage difference, and V_o is this voltage difference. In the case of non-monotonic functions a different value of R_o/V_o must be taken for each monotonic section. V_n'/V_n is the largest value of V_m'/V_m for any value of m (Thus $S_n = \infty$).

The effect of the resistors S_m is to cause the measured voltages across the segment to approximate more closely to V_m'.

Application of the Method

When the errors are only of types (2) and (4) the method is exact, i.e. with the extra shunts S_m, the measured voltages will be V_m' exactly.

When the errors of types (1) and (3) are not large (as will be the case in most analogue computer applications),

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the method will give a substantial improvement in the accuracy of the function.

The nature of the approximations involved in the method can be appreciated by considering the method of calculation of the resistors S_m.

Suppose a function is required in which the voltage across the m^{th} segment is V_m' and that, due to various errors, with the actual loaded potentiometer it is V_m (i.e. notation as above). Let R_m be the total segment impedence of a similar, but hypothetical, potentiometer which has the same segment voltage differences V_m, but is such that there are no effects present due to arm loading or incorrect position of the taps.

The shunts S_m are calculated to be such, that they will, when connected to this ideal potentiometer, change its segment voltages from V_m to V_m'. The assumption is made that these same shunts will have the same effect on the actual potentiometer. When the electrical differences between the actual and hypothetical potentiometer are small the errors involved in making this assumption will be very small. The shunts S_m are themselves only small

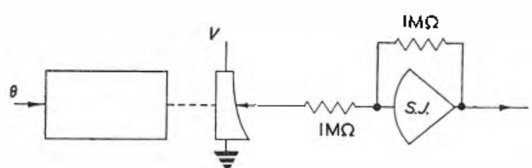

Fig. 3. Function generator potentiometer in an analogue computer circuit

correcting conductances. The difference of their effects on the two potentiometers is the only error involved.

Analysis of the Case of the Ideal Potentiometer

The function is assumed to be monotonic. An extension to a general type of function is made later.

The following notation is used:

R_m = total impedance of the m^{th} segment before correction with the shunts S_m.

V_m = measured voltage across the m^{th} segment, before correction.

V_m' = required voltage across the m^{th} segment.

δV_m = change in V_m required to make the function correct (i.e. $\delta V_m = V_m' - V_m$).

δR_m = change in R_m required to make the function correct.

S_m = shunt that will change R_m to $R_m + \delta R_m$.

δI = change in potentiometer current due to the shunts S_m.

I = potentiometer current before correction with the shunts S_m.

V = constant supply voltage across the two ends of the potentiometer.

V_o = largest voltage difference across a segment (before correction).

R_o = impedance of this section.

By definition and Ohm's law:

$$V_m = IR_m \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$\Sigma V_m = V \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Taking finite differences in equation (1):

$$\delta V_m = I\delta R_m + R_m \delta I + \delta I \delta R_m \quad \ldots\ldots\ldots\ldots \quad (3)$$

Similarly with equation (2):

$$\Sigma \delta V_m = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

From equations (4) and (3):

$$\delta I \Sigma (R_m + \delta R_m) = -I \Sigma \delta R_m \quad \ldots\ldots\ldots\ldots \quad (5)$$

From equations (3) and (5):

$$\delta V_m = I\left\{\delta R_m - \frac{(R_m + \delta R_m)\Sigma\delta R_m}{\Sigma(R_m + \delta R_m)}\right\} \quad \ldots\ldots\ldots \quad (6)$$

This is an equation giving the necessary corrections in R_m, namely δR_m, that will make the function correct. It is possible to choose the value of one of the δR_m.

Let $\delta R_n = 0$ (as will be seen later, the value of n is determined, and is not arbitrary).

Then:

$$\delta V_n = -IR_n \frac{\Sigma \delta R_m}{\Sigma(R_m + \delta R_m)} \quad \ldots\ldots\ldots\ldots \quad (7)$$

Substituting equation (7) into equation (6):

$$\delta V_m = I\{\delta R_m (1 + \delta V_n/IR_n) + R_m \delta V_n/IR_n\}$$

$$= \delta R_m (I + \delta V_n/R_n) + R_m \delta V_n/R_n$$

i.e. $$\delta R_m = \frac{\delta V_m - R_m \delta V_n/R_n}{I + \delta V_n/R_n}$$

Since R_m can only be decreased by an extra shunt resistor there is a requirement that δR_m must always be negative.

Thus n must be chosen so that:

δV_m is never $> (R_m/R_n)\delta V_n$

i.e. $\delta V_m/R_m \leqslant \delta V_n/R_n$ for all m

or using equation (1):

$\delta V_m/V_m \leqslant \delta V_n/V_n$

This equation defines the n^{th} segment as the one requiring the biggest percentage increase in potential difference.

By definition:

$$\delta R_m = \frac{R_m S_m}{R_m + S_m} - R_m = \frac{-R_m^2}{R_m + S_m}$$

(S_m is the correcting shunt resistor which changes R_m to $R_m + \delta R_m$).

i.e. $$S_m = (-R_m^2/\delta R_m) - R_m$$

$$= -R_m^2 \frac{(I + \delta V_n/R_n)}{\delta V_m - (R_m/R_n)\delta V_n} - R_m$$

$$= R_m \frac{(1 + \delta V_n/V_n)}{\delta V_n/V_n - \delta V_m/V_m} - R_m$$

$$= \frac{R_m(1 + \delta V_m/V_m)}{\delta V_n/V_n - \delta V_m/V_m} = \frac{R_m V_m'}{V_m(V_n'/V_n - V_m'/V_m)}$$

(Since $V_m' = V_m + \delta V_m$, $V_n' = V_n + \delta V_n$)

Now $R_m = V_m/I$

and

$$R_o = V_o/I \qquad R_m/V_m = R_o/V_o$$

$$\therefore S_m = \frac{R_o V_m'}{V_o(V_n'/V_n - V_m'/V_m)}$$

When the function is not monotonic the same formula applies, but for each monotonic section of the curve, R_o/V_o will have a different value (it is the reciprocal of the current in each section).

Practical Examples

Two examples of actual applications of the method are given.

In the first, the arm loading is relatively small (1MΩ

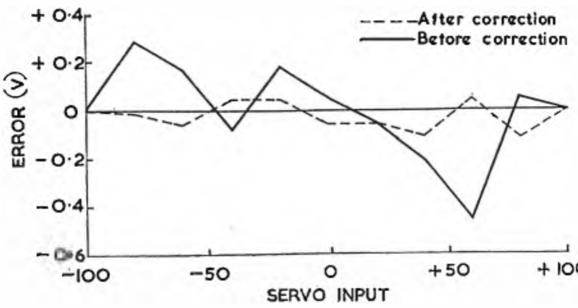

Fig. 4. Example 1, function error before and after correction

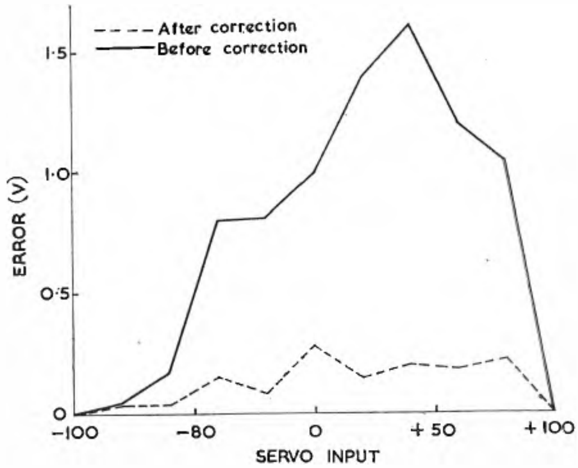

Fig. 5. Example 2, function error before and after correction

load on a 28kΩ function potentiometer). The function errors are thus mainly due to deviations of the potentiometer loading resistors from the design value, r_m. In Fig. 4 the function errors are plotted before and after correction. The accuracy of measurement of the function was of the same order as the indicated residual errors in the function after correction.

In the second example the function potentiometer was loaded with a set of resistors and the function measured with a 1MΩ arm load. This measurement was then taken as the design objective to be achieved with an arm loading of 100kΩ. In this case, therefore, most of the function error was due to arm loading. The function errors before and after correction are shown in Fig. 5. In both Figs. 4 and 5 the errors shown are those in the actual value of the function at any point, and not the errors in the segment potential differences.

In both examples, errors due to incorrect positions of the potentiometer taps are negligible. The position of the taps was found experimentally and all measurements were taken at these points.

The analogue circuit used is shown in Fig. 3 (in the second example both 1MΩ resistors were changed to 100kΩ).

The calculation of the original set of loading resistors r_m (that would give the correct function if there were no arm loading or other errors) is not shown. A simple method of determining these resistors in the case of a function potentiometer with equal unloaded segment impedance R_0, is as follows:

(1) Determine the desired value of the function at the break points (i.e. tap points) (E_m).

(2) Compute the segment potential differences $E_m - E_{m-1}$.

(3) Determine the maximum value of $E_m - E_{m-1}$ (i.e. V_0).

The segment loading resistors are given by:

$$r_m = \frac{R_0}{V_0(E_m - E_{m-1})^{-1} - 1}$$

Although resistance values have been worked out to several places of decimals, only single standard value carbon resistors were used in practice. This resulted in deviations

EXAMPLE 1

SERVO INPUT, θ (NOMINAL) (VOLTS)	FUNCTION VALUE, $f(\theta)$ (DESIGN OBJECTIVE) (VOLTS)	V_m' (VOLTS)	FUNCTION VALUE (MEASURED BEFORE CORRECTION) (VOLTS)	V_m (VOLTS)	$\frac{V_m'}{V_m}$	$\frac{V_n'}{V_n} - \frac{V_m'}{V_m}$	$\frac{R_0 V_m'}{V_0}\left(\frac{V_n'}{V_n} - \frac{V_m'}{V_m}\right)^{-1}$ (k Ω)	FUNCTION VALUE (MEASURED AFTER CORRECTION) (VOLTS)
−100	0		0					0
		6		6·29	·95390	·07070	10·40	
−80	6		6·29					5·99
		9		8·88	1·01351	·01108	99·57	
−60	15		15·17					14·95
		14		13·75	1·01818	·00641	267·7	
−40	29		28·92					29·05
		3		3·25	0·92308	·10151	3·62	
−20	32		32·17					32·05
		5		4·88(n)	1·02459	0	∞	
0	37		37·05					36·94
		16		15·90	1·00629	·01830	107·1	
20	53		52·95					52·95
		23		22·85(V_0)	1·00656	·01803	156·3	
40	76		75·80					75·90
		12		11·75	1·02128	·00331	443·8	
60	88		87·55					88·05
		3		3·5	0·85714	·16745	2·195	
80	91		91·05					90·89
		9		8·95	1·00559	·01900	58·03	
100	100		100					100

Function potentiometer impedance 28$k\Omega$
Impedance (R_0) of segment with largest voltage difference (V_0) 2·8$k\Omega$
Arm loading 1 MΩ

EXAMPLE 2

SERVO INPUT, θ (NOMINAL) (VOLTS)	FUNCTION VALUE, $f(\theta)$ (DESIGN OBJECTIVE) (VOLTS)	V_m' (VOLTS)	FUNCTION VALUE (MEASURED BEFORE CORRECTION) (VOLTS)	V_m (VOLTS)	$\frac{V_m'}{V_m}$	$\frac{V_n'}{V_n} - \frac{V_m'}{V_m}$	$\frac{R_0 V_m'}{V_0}\left(\frac{V_n'}{V_n} - \frac{V_m'}{V_m}\right)^{-1}$ (k Ω)	FUNCTION VALUE (MEASURED AFTER CORRECTION) (VOLTS)
−100	0		0					
		6·28		6·23	1·00803	·03859	20·17	
−80	6·28		6·23					6·25
		8·90		8·77	1·01482	·03180	34·69	
−60	15·18		15·00					15·15
		13·92		13·30(n)	1·04662	0	∞	
−40	29·10		28·3					28·95
		3·17		3·15	1·00635	·04027	9·757	
−20	32·27		31·45					32·19
		4·85		4·67	1·03854	·00808	74·40	
0	37·12		36·12					36·85
		15·88		15·48	1·02584	·02078	94·70	
20	53·00		51·6					52·85
		22·80		22·59(V_0)	1·00930	·03732	75·72	
40	75·80		74·19					75·60
		11·97		12·38	0·96688	·07974	18·61	
60	87·77		86·57					87·60
		3·32		3·46	0·95954	·08708	4·73	
80	91·09		90·03					90·85
		8·91		9·97	0·89368	·15294	7·22	
100	100		100					100

Function potentiometer impedance 28kΩ
Impedance (R_0) of segment with largest voltage difference (V_0) 2·8kΩ
Arm loading 100kΩ

from the nominal value of up to about 5 per cent of both the original loading resistors r_m and the correcting set S_m. It is an essential feature of this method that fairly wide tolerance resistors may be used throughout. This is because the resistors S_m that correct the errors in the r_m make only a small contribution to the total conductance of each segment. Thus they themselves do not have to be very accurate.

Acknowledgment

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The Generation of Fourier Transforms and Coefficients on an Analogue Computer

By F. C. Harbert*

Two computer arrangements are described for obtaining the Fourier transform of a function of time; one requiring computing amplifiers and multipliers, the other computing amplifiers only. A method is described for obtaining on an analogue computer the frequency response of a physical system once its response to a step input is known. This technique is illustrated by an example. An extension of the computer arrangements permits the extraction of Fourier coefficients from a periodic waveform.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

ONE of the many design aspects of a new servo-mechanical control system is the investigation of the frequency response of the proposed system before closing the loop. The characteristics of the closed loop system can then be predicted and appropriate changes can be made to avoid damaging the equipment when the loop is closed. The frequency response uniquely describes a linear physical system and the basic method for obtaining such a response is well known. A system is uniquely defined however by its complete response to any known forcing function and in particular a step impulse. The application of the Fourier transform makes possible the extraction of the frequency response of the system from this response to a unit impulse. The extraction can be carried out in a straightforward manner on an analogue computer by one of two methods.

The implications of this method become evident immediately. Whereas the frequency responsing of a system requires specialized equipment and takes up considerable time besides causing wear to the system, it is considerably easier and faster to generate a single impulse and to record the resulting transient response of the system. The analysis required to extract the frequency response from the record can be done on an analogue computer without further recourse to the system.

* *Louis Newmark Ltd.*

The definition of the Fourier transform is:

$$F(\mathrm{j}\omega) = \int_{-\infty}^{\infty} f(t)\, \mathrm{e}^{-\mathrm{j}\omega t}\, dt$$

$$= \int_{-\infty}^{\infty} f(t) \cos \omega t\, dt \;-\; \mathrm{j}\int_{-\infty}^{\infty} f(t) \sin \omega t\, dt \quad \ldots\ldots \quad (1)$$

Letting

$$A(\omega) = \int_{-\infty}^{\infty} f(t) \cos \omega t\, dt \text{ and } B(\omega) = \int_{-\infty}^{\infty} f(t) \sin \omega t\, dt \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$F(\mathrm{j}\omega) = A(\omega) - \mathrm{j}B(\omega)$$
$$= \surd\{[A(\omega)]^2 + [B(\omega)]^2\} \angle -\tan^{-1}(B(\omega)/A(\omega)) \quad \ldots\ldots\ldots\ldots \quad (3)$$

For the purpose of considering practical systems the function $f(t)$ can be restricted as follows:

$f(t) = 0$ when $t < 0$

and

$f(t) \to 0$ as $t \to \infty$

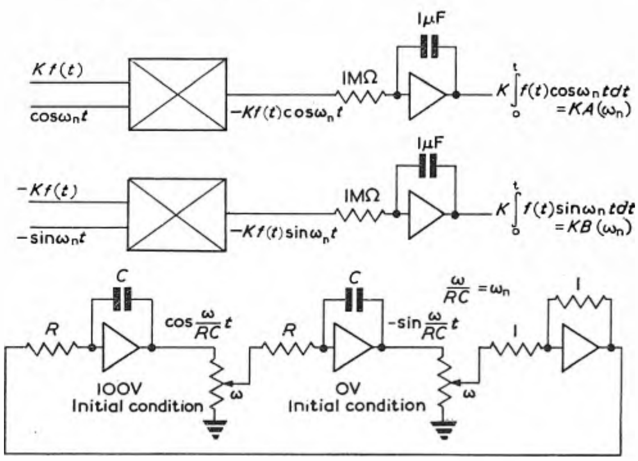

Fig. 1. Analogue computer arrangement for the generation of Fourier transforms

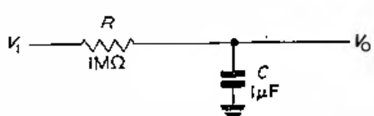

Fig. 2. *RC* network used in illustrative example

Because of these restrictions equations (2) reduce to

$$A(\omega) = \int_{0}^{\infty} f(t) \cos \omega t\, dt, \quad B(\omega) = \int_{0}^{\infty} f(t) \sin \omega t\, dt \quad \ldots\ldots\ldots\ldots \quad (4)$$

For any selected value of ω the two integrations above can be performed by the circuit in Fig. 1 which requires a function multiplier, a sinusoidal generator operating at a circular frequency of ω and an integrating amplifier.

As the function $f(t)$ approaches zero the output of the integrators will approach constant values which represent $A(\omega)$ and $B(\omega)$, and the process may be repeated with selected values of ω to cover the frequency spectrum required. When the values cf $A(\omega)$ and $B(\omega)$ are known for a sufficient number of discrete frequencies a polar plot may be drawn. As indicated in equation (3) the transform is a vector having the magnitude $\surd\{[A(\omega)]^2 + [B(\omega)]^2\}$ and phase-angle $-\tan^{-1}(B(\omega)/A(\omega))$ which completely describe the Fourier transform of the function $f(t)$.

To clarify the technique described above consider the *RC* network in Fig. 2. It will be supposed that this represents a system of which the frequency response is required. This will be obtained by using the Fourier transform.

The first step is to apply a unit impulse and record the output as a function of time. The recording can be made

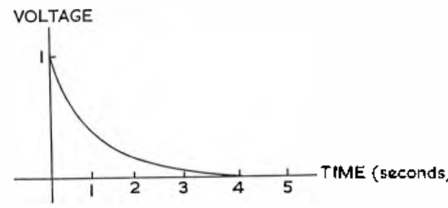

Fig. 3. Time response of *RC* network to unit impulse

on tape, pen-recorder or with a camera and oscilloscope or alternatively fed direct into the analogue computer. It is known that the response of this network to a step impulse is given by $f(t) = \mathrm{e}^{-t}$ and the response which should be obtained is shown graphically in Fig. 3.

The next step is to obtain the Fourier transform of the characteristic time function $f(t)$ using an analogue computer to perform the necessary mathematical operations of multiplications and time integration. This example, however, will be proceeded with analytically.

$$F(\mathrm{j}\omega) = \int_{0}^{\infty} \mathrm{e}^{-t}\, \mathrm{e}^{-\mathrm{j}\omega t}\, dt \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (5)$$

$$= \int_{0}^{\infty} \mathrm{e}^{-t(1+\mathrm{j}\omega)}\, dt$$

$$= -\left[\frac{\mathrm{e}^{-t(1+\mathrm{j}\omega)}}{1+\mathrm{j}\omega}\right]_0^\infty$$

$$= \frac{1}{1+\mathrm{j}\omega} = \frac{1-\mathrm{j}\omega}{1+\omega^2} \quad \ldots\ldots\ldots\ldots \quad (6)$$

Whence $A(\omega)\ \dfrac{1}{1+\omega^2}$

$$B(\omega) = \frac{\omega}{1+\omega^2} \quad \ldots\ldots\ldots\ldots \quad (7)$$

The result of equation (6) checks with the transfer function of the *RC* network and could have been obtained by circuit theory. If the computer had been used to perform the integration two separate outputs would have been furnished; $A(\omega)$ and $B(\omega)$ the real and imaginary parts of

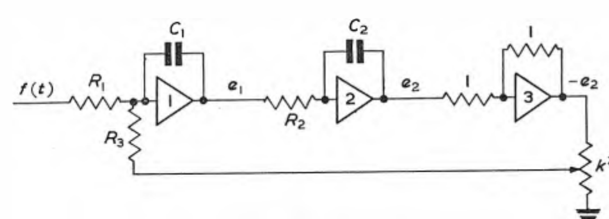

Fig. 4. Alternative computer arrangement for the generation of Fourier transforms

the above transfer function from which a polar plot (i.e. Nyquist diagram) could have been made directly.

Alternative Method

The analogue computer can be used to find the Fourier transform of physical systems without requiring explicit use of multipliers. It will now be shown that by using the circuit in Fig. 4 the voltage e_2 will be a sine wave having a amplitude proportional to $\sqrt{\{[A(\omega_n)]^2+[B(\omega_n)]^2\}}$ and a phase-shift equal to $-\tan^{-1}(B(\omega_n)/A(\omega_n))$ where ω_n is the undamped natural frequency of the circuit controlled by the potentiometer. By setting ω_n to a number of different values in succession over the frequency range of consequence, the amplitude and phase-shift of the vector $F(j\omega)$ may be determined and plotted. The same restrictions on $f(t)$ must be observed in this method. These restrictions do not handicap the analysis of practical systems.

In Fig. 4 let the output of amplifier (1) be e_1; the output of amplifier (2) be e_2: relating these outputs to the inputs:

$$e_1 = -R_2C_2se_2$$

$$-e_1 = (I/R_1C_1)\int_0^t f(t)\,dt - (k^2/R_3C_1)\int_0^t e_2\,dt \quad \ldots\ldots (8)$$

Whence by differentiation:

$$s^2e_2 + \frac{k^2}{R_2R_3C_1C_2}\,e_2 = \frac{f(t)}{R_1R_2C_1C_2} \quad \ldots\ldots (9)$$

where $e_1 = e_2 = 0$ when $t = 0$

Let $(k^2/R_2R_3C_1C_2) = \omega_n^2$

Equation (9) then becomes:

$$s^2e_2 + \omega_n^2e_2 = \frac{f(t)}{R_1R_2C_1C_2} \quad \ldots\ldots (10)$$

Taking the Laplace transform of this equation:

$$E_2(s) = \frac{F(s)/R_1R_2C_1C_2}{s^2+\omega_n^2} \quad \ldots\ldots (11)$$

Now noting that:

$$L^{-1}\left[\frac{1}{s^2+\omega^2}\right] = 1/\omega_n\,\frac{\sin\omega_n t}{\omega_n}$$

one may solve the equation for the general case by using the superposition theorem which states that:

$$L^{-1}\left[\frac{F(s)/R_1R_2C_1C_2}{s^2+\omega_n^2}\right] =$$

$$\frac{1}{\omega_nR_1R_2C_1C_2}\int_0^t f(\tau)\sin\omega_n(t-\tau)\,d\tau \quad \ldots\ldots (12)$$

Algebraic manipulation reduces this to:

$$e_2 = \frac{1}{\omega_nR_1R_2C_1C_2}\left[\sin\omega_n t\int_0^t f(\tau)\cos\omega_n\tau\,\delta\tau - \cos\omega_n t\int_0^t f(\tau)\sin\omega_n\tau\,d\tau\right] \quad \ldots\ldots (13)$$

It can be seen from equation (13) that if the integration is extended far enough in time the integrals approach $A(\omega)$ and $B(\omega)$ as defined in equation (3), so that:

$$e_2 = \frac{1}{\omega_nR_1R_2C_1C_2}[A(\omega_n)\sin\omega_n t - B(\omega_n)\cos\omega_n t]$$

$$= \frac{1}{\omega_nR_1R_2C_1C_2}\sqrt{\{[A(\omega_n)]^2+[B(\omega_n)]^2\}}\,\sin(\omega_n t+\phi)$$

where

$$\phi = -\tan^{-1}(B(\omega_n)/A(\omega_n))$$

It is important to note that the steady state values of e_2 in the computer solution are not reached until $f(t)$ has become zero. It is therefore necessary to generate an entirely independent sine wave, having the same natural frequency as the circuit in Fig. 4, to serve as a clock for measurement of phase-shift. This can readily be done with the same type of generating circuit, using three computing amplifiers where an initial condition is applied to the first amplifier instead of the forcing function. The output sine wave from the circuit of Fig. 4 and from the independent generator can then be compared after a suitable period of time to determine the phase relationship.

Since in Fig. 4 the output of amplifier (1) e_1 is tied by a unique relationship to e_2, either amplifier output may be used to determine the required Fourier transform.

Fourier Coefficients

A slight modification to the preceding technique yields the Fourier coefficients of a periodic waveform provided of course that one complete period is available. An identical computer arrangement is used but the natural frequency ω_n of the computer loop is adjusted so that n complete cycles occur during time T, i.e. $\omega_n = 2\pi n/T$ and the solution is terminated at the end of the period T so that the outputs of amplifiers (1) and (2) can be measured.

It can be shown that:

$$e_2 = \frac{-Tb_n}{2\omega_n R_1R_2C_1C_2} \qquad e_1 = \frac{-Ta_n}{2R_1C_1} \quad \ldots (14)$$

where a_nb_n are the Fourier coefficients defined as follows:

$$a_n = 2/T\int_0^T f(t)\,\cos(2\pi n/T)\,t\,dt \quad \ldots\ldots (15)$$

$$b_n = 2/T\int_0^T f(t)\,\sin(2\pi n/T)\,t\,dt \quad \ldots\ldots (16)$$

The proof can be derived from equation (12) in the preceding section. The natural frequency ω_n is changed to $2\pi n/T$ and the integration is now between 0 and T. Then equation (12) becomes:

$$e_2 = \frac{T}{2\pi n\,R_1R_2C_1C_2}\int_0^T f(t)\,\sin(2\pi n/T)(T-t)\,dt \quad \ldots (17)$$

$$e_2 = \frac{T}{2\pi n\,R_1R_2C_1C_2}\left[\sin(2\pi nT/T)\int_0^T f(t)\cos(2\pi nt/T)\,dt - \cos(2\pi nT/T)\int_0^T f(t)\sin(2\pi nt/T)\,dt\right] \quad \ldots\ldots (18)$$

$$= -\frac{T}{2\pi n\,R_1R_2C_1C_2}\int_0^T f(t)\,\sin(2\pi nt/T)\,dt \quad \ldots\ldots (19)$$

$$= -\frac{T^2b_n}{4\pi n\,R_1R_2C_1C_2} \quad \ldots\ldots (20)$$

from the definitions given in equations (16).

The relationship for e_1 can be found in a similar way.

Thus the computer circuit operates as an harmonic analyser and as different values of n are used the Fourier coefficients are available so that the complete Fourier series may be obtained.

The use of a d.c. tape recorder facilitates the carrying out of the above computer techniques. The response of a system to a unit impulse for example can be recorded and thus stored. Similarly, when a Fourier analysis of waveforms is being performed the periodic waveform can be recorded and stored for use as difference values of n are employed.

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The Role of Space Charge Waves in Modern Microwave Devices

(Part 2)

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(Voir page 460 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 465)

Résumé of the Treatment of Space Charge Waves on Infinitely Large Beams

The basic principles of space charge waves are simply shown in the example already treated in Part 1, viz. the case of waves on a beam of infinite extent, possessing a uniform current density. This illustrates the general form of the waves and it is now possible to extend the analysis to the more practical case of a finite size beam in a metal drift tube.

It should be remembered that two important assumptions were made at the outset of the study. Firstly that any transverse motion of the electrons is prevented by a very strong magnetic field. Secondly that the space charge in the unmodulated electron beam is completely neutralized by the presence of positive ions. If the beam is disturbed in some manner by a fairly high frequency signal it is then assumed that the change in velocity of the ions may be neglected when compared with the change in velocity of the electrons as the mass of the former is many times that of the latter. The wave produced by the disturbance can then be assumed to be due to the change in space charge of the electrons. These assumptions imply little practical restriction and will be retained in all subsequent discussion.

The simplest case to consider is that of an electron beam with a small sinusoidal variation in velocity added to the original d.c. velocity. This effect may be easily produced in practice by a pair of grids placed fairly close together at some point along the length of the beam. A sinusoidal voltage applied to the grids will produce sinusoidal variations in the electron beam velocity provided that the voltage applied to the grids is small compared with that used to accelerate the beam to its original velocity. Electrons leaving the grid will then have velocities which are slightly greater or smaller than that of the original beam, causing the disturbance produced to propagate down the beam. The propagation will be controlled by four factors:

(1) The charge density in the beam. This will determine the force exerted on an electron that tries to travel faster or slower than the average velocity of beam and thus experiences space charge forces.

(2) The ratio of electric charge to mass of the electron. This determines the acceleration of the electron in any given electric field.

(3) The d.c. velocity of the beam. (The wave motion is carried along with the electron beam.)

(4) The magnitude and frequency of the applied velocity modulation.

It was found in Appendix (1) that solution of Maxwell's equations yielded two waves—one travelling faster than the electrons and the other more slowly. In most microwave valves the device used to detect, or more generally, extract power from the modulated electron beam is unable to distinguish between the two waves. For this reason much of the available literature considers the *combined* output due to the two waves. In the latest types of electron beam device one is interested in the separate waves and it is essential that the differences between the separate and the combined waves should be appreciated. As seen in Part 1, a device that launches both waves on an electron beam that originally has a velocity u_0, produces a voltage amplitude:

$$V_1(z) = (V_{10}\cos\beta_p z + jZ_\omega J_{10}\sin\beta_p z)\exp(-j\beta_e z) \quad .. \quad (18)$$

The important result of this combination of the waves is that while the slow and fast waves individually suffer only changes in phase, the combined waves beat together to produce amplitude variations. If the launching device produces waves of equal amplitude, i.e. if $V_{f0} = V_{s0}$, then the second term in equation (18) vanishes and there is a simple sinusoidal distribution of voltage. In practical cases the wavelength of the beat, $2\pi/\beta_p$, is quite long compared with that of the individual waves as $\omega_p < \omega$. In the example already considered, viz, a 1kV beam with a current density $1A/cm^2$ and $f = 10\,000$Mc/s, the ratio ω/ω_p is about 20. The beat wavelength between the slow and fast waves is then about 3·7cm, while the wavelength of the individual waves is about 0·19cm. The important differences between the individual waves and their combination may be summed up thus:

Separate Slow and Fast Waves

Only the phase varies with z. The velocities are near that of the beam itself. The phase velocities are dependent on the frequency of the applied signal.

Combined Waves

Both phase and amplitude vary with z. The wavelength of the amplitude variations is determined by the charge density of the beam and is independent of the frequency of the applied signal.

Space Charge Waves Launched by a Simple Gridded Gap

There are numerous devices that will launch space charge waves and one of the simplest to consider is a 'perfect' gridded gap. This is such that the electrons will take an infinitely small time to traverse it and if the applied voltage is sinusoidal and small compared with the beam voltage V_0, then the resulting impressed velocity variations will also be sinusoidal. In the infinite beam case the voltage and current density produced by a 'perfect' gridded gap may be deduced from the formulae of Part 1. At the gap the boundary condition on J_1 must be imposed, i.e. $J_{10}=0$. This implies that the slow and fast waves will be launched with equal voltage amplitudes since $J_{10}=(V_{f0}-V_{s0})/Z_\omega$. From equations (18) and (17) it is apparent that the voltage and current possess a simple sinusoidal variation and will be given by:

$$V_1(z) = V_{10}\cos\beta_p z\exp(-j\beta_e z) \quad .... \quad (28)$$

$$J_1(z) = j(V_{10}/Z_\omega)\sin\beta_p z\exp(-j\beta_e z) \quad .... \quad (29)$$

Most of the devices used to extract r.f. power from the beam are operated by the beam current and so the output

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device is usually placed at the point where the r.f. current is a maximum. In the case of a wave launched from a simple input gap of the type just discussed this occurs at a distance $z=\pi/2\beta_p$ which is $\frac{1}{4}$ plasma wavelength. At this point the a.c. velocity in the beam is zero so that all the electrons are travelling at the d.c. velocity, u_o. For a perfect input gap the power in the beam as it enters the output device is exactly the same as when it left the electron gun. As the beam passes through the output, the combined effect of the slow and fast waves induces a voltage in the output device which in turn tends to slow down some electrons and accelerate others. If the output device has a finite resistance this process results in a reduction in the average velocity of the electrons leaving the output so that energy is extracted from the beam and appears as r.f. power. It is important to remember that the wave launched by the input device is not amplified as it travels down the beam. Conversion of d.c. to r.f. power, with resultant amplification, only occurs when an output circuit is coupled to the beam. In certain other types of devices to be described

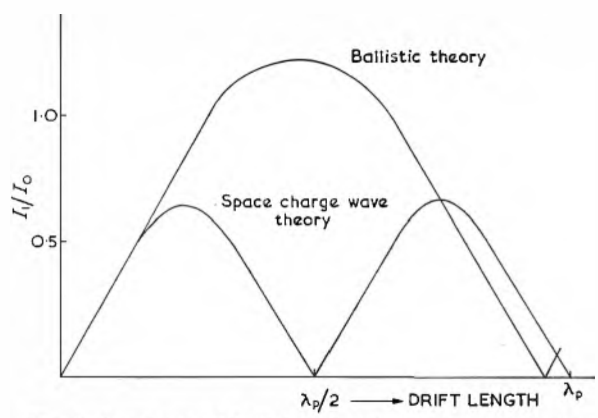

Fig. 2. Comparison of ballistic and space charge wave theories

later the amplification and conversion of power actually occurs in the beam itself.

The simple form of amplifier just described is well known in the form of the two-cavity klystron. When this device was first examined experimentally a theory due to Webster was widely, and successfully used for design purposes. This early theory ignored the effect of space charge and merely considered the kinematical process. Webster's well-known result for the r.f. beam current is:

$$I_1 = 2I_o \,.\, J_1(\beta_e z \,.\, (V_1/2V_o)) \quad \ldots\ldots\ldots \quad (30)$$

This result seems so different from that obtained by space charge wave theory that it is surprising that it gave useful results. The reason for its success can be seen by examining Fig. 2, where the results of the space charge wave theory and the kinematic theory are plotted together. It is seen that when the drift length is short compared with the plasma wavelength then the two theories give similar results. In many of the early klystrons the beam current density was quite low and consequently the plasma wavelength rather long so that the distance between the input and output cavities was very small compared with the plasma wavelength.

Before continuing to examine the effect of a finite beam size and a metal drift tube it is worth summarizing the limitations of the simple space charge wave theory as applied to the resultant of both slow and fast waves.

(1) The magnetic field must be strong enough to prevent any transverse motion of the electrons.

(2) The frequency must be such that the a.c. excursions of the ions may be neglected.

(3) The a.c. velocities imparted to the electrons by the device that launches the waves must be small compared with the beam velocity.

(4) The a.c. quantities in the beam must be small enough to justify neglecting the products of these small quantities. This assumption is made in formulating the wave equation.

The last limitation is often the most severe particularly in beams where the beam current density, and consequently space charge, is small. In the case of the simple klystron this limitation can be expressed quantitatively by the condition:

$$V_1/V_o \,.\, (\omega/2\omega_p) < 1$$

Space Charge Waves on Finite Beams, Multi-Mode Theory

The simple consideration of space charge waves on an infinitely large beam illustrates the production of the slow and fast waves and the beat that occurs between them, but is not very useful in practice. The space charge waves on a finite beam, usually in a drift tube, are of more practical interest and are similar in nature to those in the infinite beam. In the finite beam it is found that instead of one pair of waves there are an infinite number of pairs. An exact numerical computation of the combined output from all these pairs of waves is extremely tedious, but in many cases the amplitudes of many of the pairs of waves is so small that they may be neglected. The analysis was first demonstrated by Ramo[1] and Hahn[2] in 1939. As in the infinite beam case Maxwell's equations are applied to the system but there are now two boundary conditions, firstly at the junction between the electron beam edge and free space, secondly at the junction of free space with the metal drift tube.

The resulting wave equations show that the propagation constants of the space charge waves are again very nearly that of the beam itself. The difference between the constants is denoted by a small fraction δ thus:

$$\beta_s = \beta_e(1 + \delta) \quad \ldots\ldots\ldots\ldots \quad (31)$$

$$\beta_f = \beta_e(1 - \delta) \quad \ldots\ldots\ldots\ldots \quad (32)$$

In the infinite beam case it was found that there were only two waves so that δ had only one value.

In the finite beam case it is worth studying the results of Ramo's work as it illustrates how the solution yields an infinite number of values for δ.

Ramo's study of the boundary conditions for a beam of radius b inside a drift tube of radius a, yields the equation:

$$f_1(Tb) = -(Tb)\,\frac{J_1(Tb)}{J_o(Tb)} = \tau b\,\frac{K_o(\tau a)\,I_1(\tau b) - I_o(\tau a)\,K_1(\tau b)}{K_1(\tau a)\,I_o(\tau b) - I_1(\tau a)\,K_o(\tau b)} \quad \ldots\ldots\ldots\ldots \quad (33)$$

where:

$$\tau = \sqrt{(\beta^2 - k^2)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (34)$$

and:

$$T^2 = (\beta^2 - k^2)\,((\omega_p/\delta\omega)^2 - 1) \quad \ldots\ldots \quad (35)$$

In nearly all cases the beam velocity is such that $\beta^2 \gg k^2$ and $\beta \simeq \beta_e$, and hence from equation (35) it is seen that:

$$\delta = \frac{\omega_p}{\sqrt{[\omega^2 + (Tu_o)^2]}} \quad \ldots\ldots\ldots\ldots \quad (36)$$

The propagation constants may thus be found if T can be determined from equation (33).

Ramo's method of solution is to write $\tau \simeq \beta \simeq \beta_e$ so that the r.h.s. of the equation can be evaluated for the particular values of a and b used. The l.h.s., $f_1(Tb)$ is then evaluated for a range of values of T and plotted, the form of the variation being shown in Fig. 3. The intersections of this curve with the value obtained from the r.h.s. define an infinite number of values of T, which, in this approximate analysis, are independent of the beam current density. It is now possible to find a series of values of δ for any given beam current density. If the solutions of T are denoted by $T_1, T_2, T_3 \ldots T_n$, then those for δ are $\delta_1, \delta_2, \delta_3 \ldots \delta_n$. There are now an indefinite numer of pairs of space charge waves having propagation constants $\beta_1, \beta_2, \beta_3 \ldots \beta_n$. The amplitudes of these pairs may be found to be proportional to:

$$\frac{1}{(T_n b)^2\,(1 + (T_n b / K_n))}\; 1/\delta_n$$

where:

$$K_n = -T_n b\,\frac{J_1(T_n b)}{J_0(T_n b)} \quad \ldots\ldots\ldots\ldots (37)$$

The mode corresponding to the lowest value of T_n, i.e. $n = 1$, has the largest amplitude. The amplitudes of the

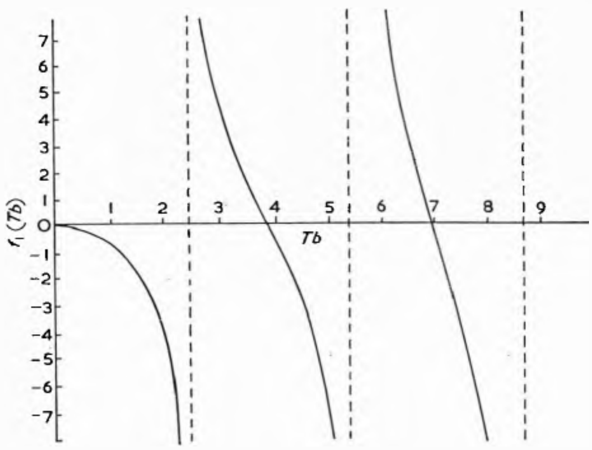

Fig. 3. Relationship between Ramo's function f_1 (Tb) and Tb

higher order modes fall off as the mode number increases. The relative amplitudes of successive modes depend on a, b, and β_e. In certain cases the first mode is predominant and the higher order modes are small enough to be neglected. The propagation constant for the *first* mode is given by:

$$\beta_1 = \beta_e(1 \pm \delta_1) = \beta_e \pm \frac{\beta_p}{\sqrt{[1 + (T_1/\beta_e)^2]}} \quad \ldots\ldots (38)$$

It is interesting to note that this expression is similar to that obtained in the case of an infinite beam except that β_p is now replaced by $\beta_p/\sqrt{[1 + (T_1/\beta_e)^2]}$. The latter quantity ($\times u_0$) is called the reduced plasma radian frequency ω_q and the factor $[1 + (T_1/\beta_e)^2]^{-\frac{1}{2}}$ the plasma reduction factor. As the reduction factor depends only on a, b, and β_e it may be plotted for a wide range of values of these quantities and the results used in klystron calculations. The wavelengths associated with ω_p and ω_q are λ_p and λ_q. The maximum r.f. current in the *first* pair of waves in the finite beam occurs when $z = (\lambda_q/4)$.

The higher order modes all have wavelengths that are greater than that of the first mode but these wavelengths are not related in any simple way. If the higher order modes are not negligible then the total r.f. current in the beam at any point can only be found by finding the contribution from each mode. The resulting variation in the r.f. current with length will not be sinusoidal even though the distribution of individual modes is sinusoidal.

Another important feature of the finite beam case is that the r.f. current is no longer evenly distributed over the whole cross-section of the beam. The radial distribution is different for each mode and is dependent on $J_0(T_n r)$ where r is the radial position in the beam. Once again the actual current at any point has to be found by adding the currents in all the modes. The general effect of the radial distribution is that more of the r.f. current is carried in the outer part of the beam than the inner.

The differences between the space charge waves in finite and infinite beams may be summed up thus.

Infinite Beam	*Finite Beam and Drift Tube*
One pair of waves.	Infinite number of pairs of waves.
Sinusoidal variation of r.f. current with z.	Individual wave pairs have sinusoidal distribution of r.f. current with z but total current distribution not sinusoidal.
No radial variation of r.f. current.	Radial variation of r.f. current.
Maximum of r.f. current at $\frac{1}{4}$ plasma wavelength	Maximum of r.f. current generally at less than $\frac{1}{4}$ wavelength.

The finite beam case is limited by the same restrictions (mentioned in the last section) as the infinite beam.

Amplitudes of the Higher Order Modes

It has already been seen that in cases where the higher order modes are negligible a modified form of the simple theory may be used where ω_p is replaced by ω_q. In cases where the higher modes are not negligible the complete theory must be used and all the significant modes included. As the amplitude of the mode always decreases with the mode order it will only be necessary to consider a finite number of the modes and it is important to know what factors affect the relative amplitudes. The expression for the r.f. current in the beam is:

$$I_1(z) = 4j\alpha I_0 \exp(-j\beta_e z) \sum_{n=1}^{\infty}\left[\frac{1}{(T_n b)^2\,(1 + (T_n b/K_n))} \times (1/\delta_n)\sin(\beta_e \delta_n z)\right] \quad \ldots\ldots\ldots\ldots (39)$$

(Where α denotes modulation depth $V_1/2V$)
This may be rewritten in the form:

$$I_1(z) = 4j\alpha I_0 \sum_{n=1}^{\infty} A_n \sin \beta_e \delta_n z \,.\, \exp(-j\beta_e z) \quad \ldots . (40)$$

where:

$$A_n = 1/\delta_n \left(\frac{1}{(T_n b)^2(1 + (T_n b/K_n))}\right)$$

and

$$\delta_n = \omega_p/\omega\,\frac{1}{\sqrt{[1 + (T_n/\beta_e)^2]}} \quad \ldots\ldots\ldots (41)$$

Thus the relative amplitudes of the modes are determined by:

$$\frac{(1 + (T_n/\beta_e)^2)^{\frac{1}{2}}}{(T_n b)^2(1 + (T_n b/K_n))} \quad \ldots\ldots\ldots\ldots (42)$$

and this factor is only dependent upon the normalized radii of the beam and drift tube and is fixed for a tube of given dimensions and beam voltage. However, the complex relationship between the values of T_n and the tube parameters, prevent any simple analysis of the relative mode amplitudes, and particular cases must be considered.

The simplest case is that of a beam filling the drift tube. Here the value of $T_n b$ is simply found from Fig. 3 as $f_1(Tb)$ is infinity and so $T_1 b = 2{\cdot}4$, $T_2 b = 5{\cdot}5$, $T_3 b = 8{\cdot}6$, etc. Examination of equa-

TABLE 1
Effect of drift tube size on the relative amplitudes of the higher order modes

Mode \ a/b	1·0	1·2	1·5	2·0
1	100	100	100	100
2	33·5	26·5	14	12·8
3	20·5	10·1	4·9	3·6
4	16·5	4·8	1·6	1·3

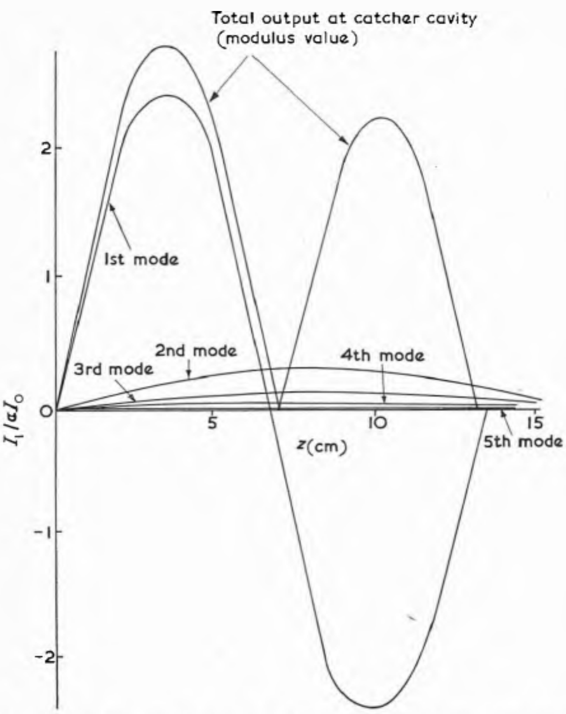

Fig. 4. Variation in amplitude of the first five space charge modes with z

tion (42) shows that the amplitudes of the modes fall off slowly if $T_n^2 \gg \beta_e^2$ and this condition will be realized if either T_n is large (i.e. b is small) or if β_e is small. It is seen that the greatest value that the second mode can have with respect to the first is $2{\cdot}4/5{\cdot}5 = 0{\cdot}43$.

In the case of a beam not filling the drift tube the calculation is a little more complicated and can only be made for one particular set of conditions. Table 1 shows the relative amplitudes of the first five modes for various different ratios of beam to drift tube size. It is seen that for the particular value of β_e chosen there is very little difference between $R = 2{\cdot}0$ and $R = \infty$. Fig. 4 shows the first five modes plotted in terms of their r.f. current contribution as a function of the drift length. The total r.f. current is also shown and it is seen that this differs appreciably from the first mode alone. The radial distribution is also affected by the ratio of the T_n's and this is illustrated for one particular case by Fig. 5. It must be remembered that all these calculations have been made for a case where the input gap is gridded so that the velocity distribution at the input gap is independent of the radius.

Space Charge Waves Launched by Gridless Gaps

If the input gap is gridded then the velocity imparted to an electron by a voltage applied to the gap is independent of the radial position of the electron. However, in the individual space charge modes the velocities are a function of radius in all cases where the beam size is finite. In order to satisfy the condition at the input gap the radial distribution of velocities in all the modes must add up to give a uniform distribution of velocities at the gridded gap. This gives the relative amplitudes of the modes as discussed in the previous section. The only physical dimension of the gap that will affect the launching of the space charge wave is its length and if this is not small compared with $2\pi/\beta_e$ then the efficiency of the gap will be less than that of a perfect gap, the difference being a factor $(\sin L/2)/(L/2)$, where $L = \beta_e l$ and l = length of gap.

If the gap is not gridded then the velocity imparted to the electrons is not independent of their radial position if the normalized radius of the beam is not negligible. Once again the total sum of the velocity distribution of the various modes has to add up to the velocity distribution in the gap, which is now a function of radius, so that the

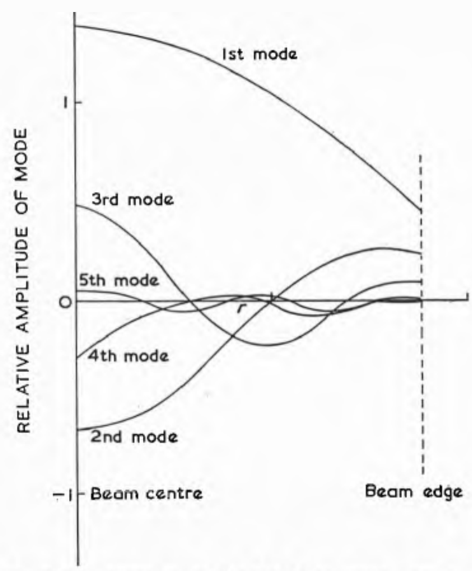

Fig. 5. Radial variation of the first five space charge modes

relative amplitudes of the various modes will be different from the gridded gap case.

Output Gap Mechanism

In any device in which fast and slow space charge waves are launched equally, no conversion of d.c. to a.c. energy actually occurs in the electron beam itself once the beam has left the device that launches the wave on it. Conversion of power only occurs when some output device is added and this often takes the form of a gridded or ungridded gap. The gap configuration will not affect the relative amplitudes of the modes in the beam (as is the case with the input gap) but the contribution made by each mode to the current appearing in the gap will depend upon the form of gap. If the gap is gridded then the coupling between the gap and each space charge mode in the beam will be the same, but for an ungridded gap, where the normalized radius of the beam is not negligible, the coupling will be different for each mode. The output current in the gap can be found by evaluating the gap factor for the radial distribution of current corresponding to each individual mode and thus finding the contribution that each mode makes to the gap current.

Practical Significance of Higher Order Modes

It is seen that an exact analysis of multi-mode space charge waves is rather tedious as it involves:

(a) Calculating the amplitude of each mode produced on the beam by the launching device.

(b) Evaluating the propagation constant for each mode.

(c) Evaluating the radial distribution of current in each mode at the output device.

(d) From the results of (b) and (c) evaluating the contribution that each mode makes to the current appearing in the output device.

Whether or not this rather lengthy procedure is necessary may be found by making a rough estimate of the relative amplitudes of the higher order modes. Unless the higher modes make a fairly substantial contribution to the total output the practical differences between an actual valve and the theoretical case considered here may make the complete analysis unjustifiable.

Small Signal Limitation of Space-Charge Wave Theory and Extensions to include Large Signals

Space charge wave theory is always designated as a 'small signal' theory but the exact limitations are often

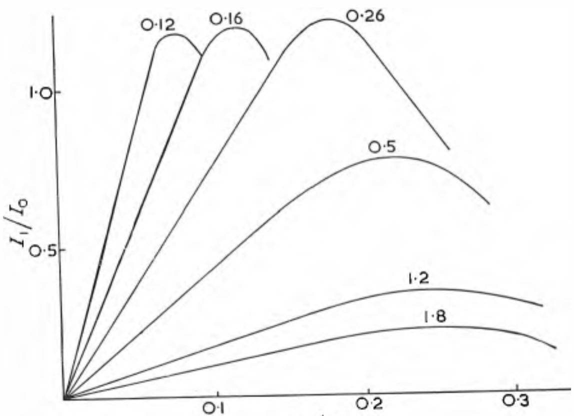

Fig. 6. Webber's large signal theory

not clearly stated. The small signal limitations can be easily defined by taking the case of an infinite beam and considering the method by which the original wave equation was obtained. In this method the products of small a.c. quantities were neglected.

e.g.:

$$\rho_1 u_1 \ll \rho_0 u_1, \quad \rho_1 u_1 \ll \rho_1 u_0$$

or:

$$\rho_1 \ll \rho_0, \qquad u_1 \ll u_0$$

The second condition is a simple one and merely states that $V_1 \ll V_0$. The conditions under which the first criterion is satisfied, in the case of a perfect gridded gap for example, can be seen by considering the final results for the magnitude of J_1, for from equation (29):

$$|J_1| = (J_0 V_1/2V_0)\ (\omega/\omega_p) \sin \beta_p z$$

whence:

$$|J_{1(\text{pk})}| = |J_0(V_1/2V_0)\ (\omega/\omega_p)|$$

Thus the a.c. component of current is comparable with the d.c. term when $(V_1/2V_0)$ (ω/ω_p) approaches unity. In many practical valves this condition is fulfilled even though V_1/V_0 is quite small and consequently this is a more stringent condition than $V_1/V_0 \ll 1$.

An exact analysis of large signal conditions in a finite beam with a drift tube has yet to be made but a recent approximate analysis by Webber[5] produces results which are extremely interesting in that they are verified by experimental observations[6]. Webber's analysis shows how the waves behave in the situation midway between pure space-charge controlled waves and ballistically controlled electron movement. The numerical results of this theory can only be obtained with the aid of a computer. It is a single mode theory and assumes that the radial distribution of r.f. current is uniform.

The computed results show that as the depth of modulation is increased, so the drift length tends to decrease. This result is similar to that obtained with simple ballistic theory but the conclusions are applicable to cases where the space charge forces, although not predominant, are by no means negligible. Fig. 6 shows how the r.f. beam current varies with the drift length for various 'normalized' depths of modulation, the significant quantity being $\omega_q V_0/\omega V_1$ rather than V_0/V_1. The effect of a finite beam is allowed for by the use of ω_q instead of ω_p, where ω_q is the modified plasma frequency.

Although Webber's theory allows for large signals, it must be remembered that if V_1 is not small compared with V_0 then the velocity variations imparted to the electrons will not be sinusoidal. This point is often overlooked and arises from the fact that the electron velocity is proportional to $\sqrt{(V_0 + V_1 \sin \omega t)}$ and the assumption is made that $V_1 \ll V_0$ so that the velocity is proportional to $V_0 + (V_1/2) \sin \omega t$. When V_1 is comparable with V_0 then the higher order terms, which have been ignored, will not be negligible.

One may sum up by saying that all theories so far quoted are only substantially correct under certain conditions. These conditions may be roughly collected together thus:

All theories: V_1/V_0 fairly small so that the velocity variations imparted to the electrons are sinusoidal.

(1) Simple ballistic theory: V_1/V_0 $(\omega/\omega_p) = (J_1/V_0) \gg 1$

(2) Space charge theory: $V_1/2V_0$ $(\omega/\omega_p) \ll 1$

(3) Webber theory: $V_1/2V_0$ (ω/ω_p) of the order of unity.

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6. Stephenson, I. M. A Two-Cavity Klystron Operating Under Large Signal Conditions. *Proc. Inst. Elect. Engrs. 107B*, 60 (1960).

(To be continued)

A MINIATURE DIRECTION FINDER

By making use of sub-miniature valves and transistors which require only small and light-weight power sources, Telefunken has designed a miniature direction finder measuring only 6¼in by 4¼in by 2¼in and weighing only slightly more than 1kg. The direction finder can be carried in the pocket or fastened to a belt and, therefore, for certain purposes its use is completely inconspicuous.

The frequency range can be divided into 10 sub-ranges between 57kc/s and 20Mc/s by the insertion of exchangeable coil cartridges. The ferrite antenna incorporated in the set is adequate for taking bearings in the near field—for a higher d.f. performance it is possible to connect a larger d.f. loop antenna which, for instance, can be fastened to the lining of the operator's jacket or coat. An auxiliary antenna for the sense determination is laid over the operator's shoulder. The frequency setting can be read from the position of the pointer on the scale whichever coil cartridge is used. The scales are calibrated in kc/s or Mc/s. The instrument for indicating the relative field strength or output is worn like a wrist-watch round the left wrist, where it can be observed without attracting undue attention. To determine the bearing direction it is merely necessary for the operator to turn his body slowly into the position of the d.f. minimum signal. If necessary, it is also possible to use the minimum volume in the headphone as a measure.

An Integrating and Differentiating Amplifier for Use in Vibration Measurements

By M. Kringlebotn*, Eng. appl. phys.

The signal source in the vibration measurements is assumed to be a velocity pick-up, and the amplifier output is delivered to a low-impedance load (e.g. a galvanometer with 25Ω internal resistance). When used as an integrator the amplifier converts the input velocity signal to a signal which indicates the displacement and, when used as a differentiator, to a signal which indicates the acceleration of the pick-up. The amplifier may also be used for linear amplification of the velocity signal. The amplifier stages are stabilized with respect to temperature changes. The additional feedback also contributes to thermal stability. A single 12·6V source is used.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

The Amplifier Circuit

The complete circuit of the amplifier is shown in Fig. 1 and is discussed briefly below.

Linear Amplification

The three amplifier stages are current controlled and parallel current feedback to stabilize the current amplication is therefore advantageous. This is accomplished by means of a feedback resistance of 10Ω at the collector side of the third stage. The feedback current is fed to the base of the first stage.

Differentiation

The input voltage is differentiated by means of a small capacitor in series with the small input resistance of the first stage. Because of the large capacitive reactance the differentiating amplifier is current controlled and therefore has the same feedback as the linear amplifier. In addition the parallel current feedback has the advantage of lowering the input resistance of the first stage.

Integration

A comparatively small internal resistance of the signal source is assumed, and the first stage of the integrating amplifier is therefore voltage controlled, while the following stages are current controlled as before. The first stage is consequently given series current feedback which stabilizes the ratio of the output current to the input voltage. This feedback is accomplished by the partly decoupled emitter resistor which also serves as the gain control for the integrating amplifier.

The output current of the first stage is integrated by means of the succeeding two stages, the first in emitter and the second in collector coupling. The capacitive feedback over these two stages is of the parallel current type and is proportional to the frequency and in opposition to the input current of the second stage. The output current is thus inversely proportional to the frequency of the input voltage as it should be for an integrating amplifier.

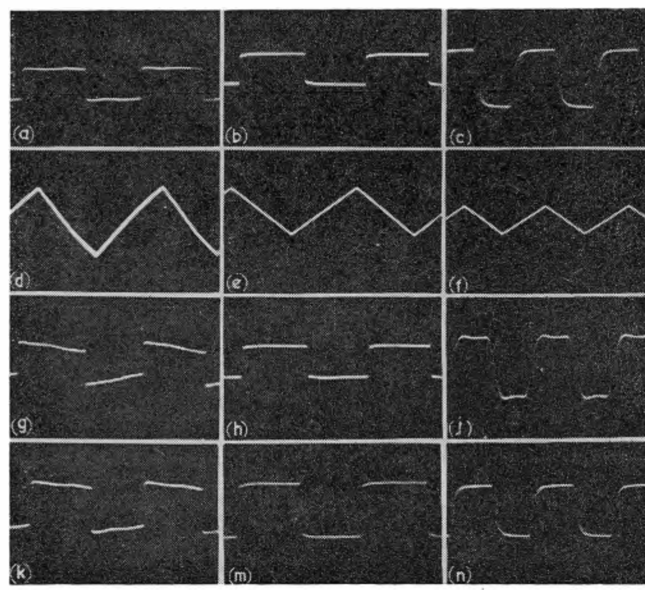
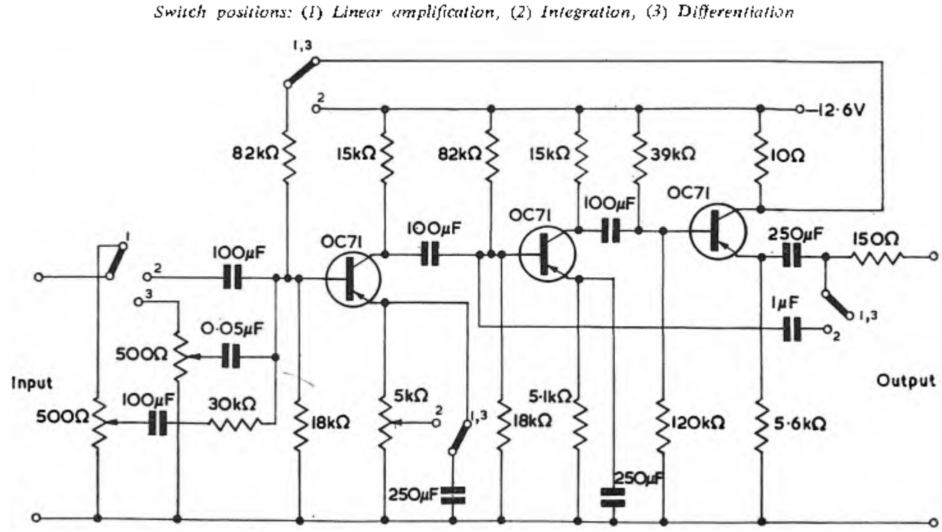


Fig. 2. The low frequency performance of the circuit

(a) (b) (c)—Input signals with frequencies of 18, 100 and 1 500c/s
(d) (e) (f)—The integrated signals
(g) (h) (j)—Differentiation of the integrated signals
(k) (m) (n)—Linear amplification of the input signals

Fig. 1. The integrating and differentiating amplifier

Switch positions: (1) Linear amplification, (2) Integration, (3) Differentiation



* *The Technical University of Norway.*

Performance

Cheap low-frequency transistors are used, but are satisfactory for the purpose. The photographs shown in Fig. 2 give a good indication of the low-frequency performance of the amplifier.

FREQUENCY-DEPENDENCE OF AMPLIFICATION

In these measurements a 25Ω resistive load was used.

Linear Amplification

The result is shown in Table 1. The maximum amplification of the circuit used as a linear amplifier (measured at 100c/s) is five times.

TABLE 1

The frequency dependence of linear amplification

FREQUENCY (c/s)	OUTPUT (mV)
Input = 20*mV*	
20	20
40	20
80	20
160	20
320	20
640	20
1280	20
2000	20
3000	20·2
4000	20·5
5000	20·8
6000	21·2
7000	21·8
8000	22·2
9000	22·9
10000	23·6

Integration

The performance as an integrator is shown in Table 2.

TABLE 2

Integration in three partly overlapping frequency ranges

FREQUENCY (c/s)	OUTPUT (mV)
Input = 20*mV*	
20	20
40	10·2
80	5·1
160	2·55
320	1·25
640	0·62
1280	0·3
Input = 157*mV*	
160	20
320	10·2
640	5·1
1280	2·5
Input = 420*mV*	
500	16
1000	8
2000	4
4000	2

At 20c/s the maximum amplification is about 11 and is obtained when there is no feedback in the first stage. In this case, however, a considerable distortion of the signal occurs.

Differentiation

The performance of the circuit when used as a differentiate is given in Table 3.

TABLE 3

Differentiation in three partly overlapping frequency ranges

FREQUENCY (c/s)	OUTPUT (mV)
Input = 2·4*mV*	
0	0·2 (noise)
20	2·6
40	5
80	10
160	20
Input = 0·3*mV*	
80	1·3
160	2·55
320	5·1
640	10·2
1280	20
Input = 0·1*mV*	
500	2·95
1000	5·8
2000	11·3
4000	20

Maximum amplification at 20c/s is about 1.

INTEGRATION FOLLOWED BY DIFFERENTIATION

Table 4 shows the result of differentiating the output from the integrator. Ideally the result should be independent of the frequency; this holds for low frequencies.

TABLE 4

Integration followed by differentiation

FREQUENCY (c/s)	OUTPUT (mV)
Input = 15*mV*	
20	16
40	16
80	16
160	16
320	16
640	16
1280	16
2000	16·2
3000	16·1
4000	15·5
5000	14
6000	12
7000	9·8
8000	8
9000	6·5
10000	5·3

Conclusion

The current drain of this circuit is only 2·3mA. The working point of the third stage is chosen to give a maximum voltage output of about 20mV to the 25Ω load.

The tables show that the circuit works satisfactorily in the frequency range of 20c/s to 3kc/s. The feedback circuits are very effective, and necessary, for the good performance of the circuit.

If a high input impedance is desired, one may add a collector coupled stage at the input.

A Novel Single Transistor RC Oscillator

By E. T. Emms*, B.Sc., A.R.C.S., A.M.I.E.E.

A novel single transistor RC oscillator is described which will oscillate with the majority of low frequency transistors even when the transistor current gain is at the lower limit. The circuit employs a modification of the parallel-T network. The effect of finite transistor input impedance and output capacitance is discussed.

(Voir page 527 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 532)

SINGLE valve *RC* oscillators have been in use for a considerable time. The stage gain requirements for valve *RC* oscillators are as follows:

Three-stage *RC* coupled -	gain required = 29
Four-stage *RC* coupled - -	gain required = 18·4
Parallel-*T* - - - -	gain required = 11

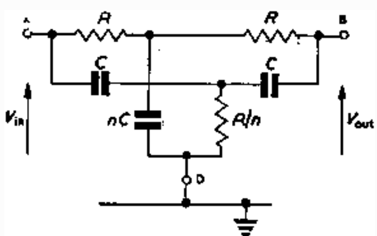


Fig. 1. Normal parallel-T network

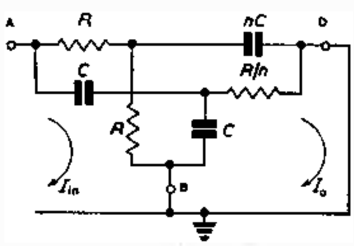


Fig. 2. Revised parallel-T network

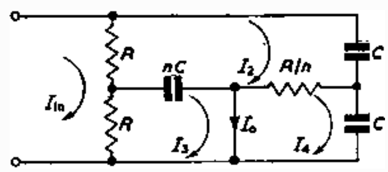


Fig. 3. Redrawn circuit of Fig. 2

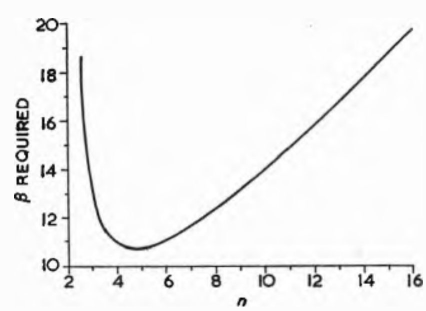


Fig. 4. Required β for values of n

It will be noted that the parallel-*T* circuit requires the least gain for maintenance of oscillation so that it is to be expected that such a circuit will reliably oscillate with valves approaching the end of their life. Alternatively advantage may be taken of this small gain requirement to design the circuit to give a very good waveform.

The only *RC* oscillators described for single transistor operation are the *RC* coupled types[1]. The required β's for these circuits are:

Three-stage *RC* coupled - -	β required > 46
Four-stage *RC* coupled - -	β required > 35

Examination of the transistor manufacturers' catalogues shows that most low frequency transistors have a lower limit of β less than these figures and it follows that circuits cannot be designed which operate reliably for all transistors of a given type. Because of this, most audio transistor oscillators described in the literature use two transistors in a Wien bridge circuit or variants of this. Accordingly it was investigated to see whether the parallel-*T* or a variant of it could be exploited to produce a reliable single transistor oscillator. The parallel-*T* circuit as normally employed is not suitable since it requires to be driven from a low impedance voltage source and be terminated in a high impedance. A variant of the circuit is shown in Fig. 2. It will be noted that terminals B and D of the normal parallel-*T* shown in Fig. 1 have been interchanged and the revised parallel-*T* network is shown working into a very small (effectively zero) impedance.

Circuit Analysis

Straightforward analysis of Fig. 3 gives:

$$\begin{vmatrix} (1+x)(1+n) & -1 & -x \\ -1 & 1+nx & 0 \\ -x & 0 & n+x \end{vmatrix} \begin{vmatrix} I_2 \\ I_3 \\ I_4 \end{vmatrix} = \begin{vmatrix} nxI_{in} \\ nxI_{in} \\ 0 \end{vmatrix}$$

where $x = pCR$

* *General Precision Systems Ltd.*

whence:

$$I_o/I_n = \frac{I_3 - I_4}{I_{in}} = \frac{xn(2+n)(1+x)}{n+(2+2n+n^2)x+(2+2n+n^2)x^2+x^3} \quad \ldots\ldots\ldots\ldots (1)$$

Putting $\omega CR = \omega/\omega_o = 1$, i.e. $x = j$

$$I_o/I_{in} = \frac{n(2+n)}{2+n+n^2} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

which indicates that at a frequency given by $\omega_o = 1/RC$ the output current is in phase with the input current to the network.

Differentiating this expression with respect to n it is found that its maximum value occurs when $n = 2(1 + \sqrt{2}) \simeq 4{\cdot}8$ and then:

$$I_o/I_{in} = \frac{2 + 4\sqrt{2}}{7} = \frac{7{\cdot}657}{7} = 1{\cdot}094$$

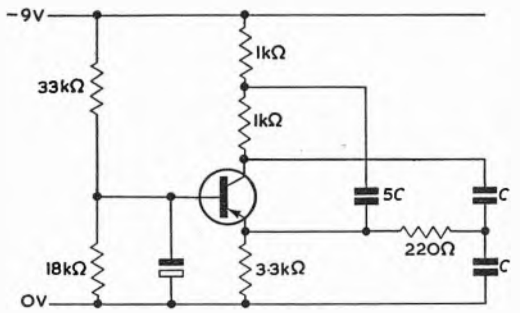


Fig. 5. Practical circuit of oscillator

Thus under these conditions the network has a current gain.

For sustained oscillations this network must be associated with a transistor stage which gives zero phase shift, input to output, and the grounded base stage satisfies this requirement. The stage gain at $\omega = \omega_o$ must be sufficient to have an overall loop gain of unity so that the transistor current gain must be equal or greater than $1/1{\cdot}094$; this implies a transistor having $\alpha > 0{\cdot}9132$ or $\beta > 10{\cdot}66$. Nearly all commercial transistors (even on lower limit) meet this requirement.

For values of n other than the optimum given above the necessary β required may be found:

$$\beta > \frac{1}{(I_o/I_{in}) - 1} = \frac{2 + n + n^2}{n - 2} \quad \ldots\ldots\ldots (3)$$

This expression is plotted in Fig. 4 and it will be noted that the value of n is not critical and therefore for design purposes one may take $n \simeq 5$.

In the above analysis it was assumed that the input resistance of the transistor was negligible compared with R. Of course the input resistance is finite, though small, and so a further analysis is given in Appendix (1) to show the effect of input resistance on the operation. At higher frequencies the collector to base capacitance becomes important and Appendix (2) discusses this effect.

Practical Circuit

Fig. 5 shows a practical realization of the oscillator. The circuit is arranged to operate at a collector current of between 1·5 and 2·0mA since this leads to the optimum β value for the transistor.

The base supply must be firmly decoupled, a rough guide being that the decoupling capacitor capacitance should be 20 to 30 times the frequency determining capacitors.

The range of frequencies obtained was large, frequencies of 15c/s to 100kc/s being obtained with ease.

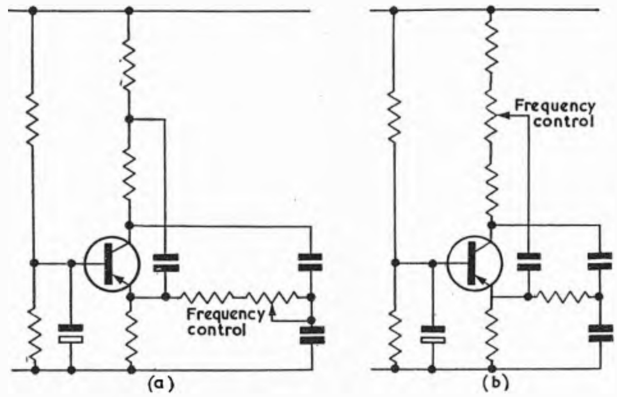


Fig. 6. Methods of frequency control

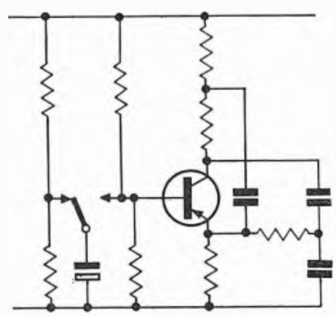


Fig. 7. Arrangement for keying the oscillator

For the lower frequencies, electrolytic capacitors were used in the frequency determining network.

With a network designed for a nominal oscillation frequency of 12·37kc/s the following results were obtained.

Supply potential (volts)	$7\frac{1}{2}$	9	$10\frac{1}{2}$
Frequency (kc/s)	12·4	12·37	12·32

Ambient Temperature (°C)	19·5	27·2	34	40
Frequency (kc/s)	12·37	12·485	12·63	12·82

No difficulty should be experienced in providing any of the standard frequency/temperature stabilizing techniques.

The circuit works equally well with pnp or npn transistors (with appropriate change of supply of course).

Fig. 6 shows two methods by means of which the frequency may be varied over a small range.

Keying the Oscillator

It is frequently desired to key an oscillator. Many circuits produced for this purpose suffer from the disadvantage that the anode current or collector current is 'turned' on or off which results in a step of current. This gives rise to objectionable 'key clicks' if the output is fed into a headset or loudspeaker. This objection may be obviated by the use of the circuit shown in Fig. 7. In this circuit, it will be seen that the decoupling capacitor on the base is removed and re-introduced by operation of the keying member (this may be a key, relay contacts or a switched transistor). While the capacitor is removed the circuit will not oscillate. The capacitor is maintained charged to its normal operating potential. By this means the d.c. conditions in the transistor oscillator do not change whether it is oscillating or quiescent and hence 'key clicks' are obviated.

Appendix (1)

Effect of Finite Input Impedance of the Transistor

If it is assumed that the input resistance of the transistor can be represented by μR then the mesh matrix is modified to:

$$\begin{vmatrix} 1+n)(1+x) & -1 & -x \\ -1 & 1+nx+\mu nx & -\mu nx \\ -x & -\mu nx & n+x+\mu nx \end{vmatrix} \begin{vmatrix} I_2 \\ I_3 \\ I_4 \end{vmatrix} = \begin{vmatrix} nxI_{in} \\ nxI_{in} \\ 0 \end{vmatrix}$$

whence:

$$I_o/I_{in} = \frac{xn(2+n)(1+x)}{n+(\phi+n\theta)x+(\phi+2n\theta)x^2+n(1+\theta)x^3}$$

where $\theta = \mu(2+n)$ $\phi = 2+2n+n^2$

Again putting $x = j\omega_0 CR = j\omega'$ the condition for zero phase shift between input and output is found to be:

$$(\omega')^2 = (\omega_0 CR)^2 = \frac{1}{1+\theta} = \frac{1}{1+\mu(2+n)}$$

so the 'resonant' frequency is:

$$\omega_0 = \frac{1/CR}{\sqrt{[1+\mu(2+n)]}}$$

The current gain at 'resonance' is then given by:

$$I_o/I_{in} = \frac{n(2+n)}{2+n+n^2+\mu n(2+n)}$$

and the necessary transistor β to maintain oscillation is:

$$\beta > \frac{2+n+n^2+\mu n(2+n)}{n-2-\mu n(2+n)}$$

Taking the normal design value of $n = 5$ these equations become:

$$\omega_0 = \frac{1/CR}{\sqrt{(1+7\mu)}}$$

and:

$$\beta > \frac{32+35\mu}{3-35\mu}$$

Table 1 may be constructed.

TABLE 1

μ	Required β	*Resonant frequency* (with μ=0)
0	10·66	1·0
0·01	12·20	0·965
0·02	14·18	0·935
0·05	26·9	0·861
0·07	62·5	0·817

It will be seen that the presence of finite input resistance lowers the 'resonant' frequency below that given by the simple equation $f = 1/2\pi RC$ and further the required β may become quite large if the input resistance becomes an appreciable fraction of the network R.

Appendix (2)

The Effect of Transistor Output Capacitance

If the transistor output capacitance be represented by C/λ then a straightforward though lengthy analysis yields:

$$I_o/I_{in} = \frac{xn(2+n)(1+x)}{n+(2+2n+n^2)x+(2+2n+n^2)x^2+nx^3+(x/\lambda)\{2n+(4+n^2x\ 2nx^2\}}$$

This expression is real at a frequency given by:

$$\omega^2C^2R^2 = \frac{2+3n+n^2+(2n/\lambda)}{2+3n+n^2+\dfrac{4+2n+n^2}{\lambda}}$$

and then:

$$I_o/I_{in} = \frac{n(2+n)\{2+3n+n^2+((4+2n+n^2)/\lambda)\}}{(2+2n+n^2+(2n/\lambda))\left(2+3n+n^2+\dfrac{4+2n+n^2}{\lambda}\right)-(2+3n+n2+(2n/\lambda))\left(\dfrac{n+2n}{\lambda}\right)}$$

Taking n as the normal value of 5 these equations reduce to:

$$\omega CR = \sqrt{\left(\frac{42+(10/\lambda)}{42+(39/\lambda)}\right)}$$

and

$$\beta = \frac{1344+(1393/\lambda)+(290/\lambda 2)}{126-(28/\lambda)-(290/\lambda 2)}$$

Table 2 may be constructed.

TABLE 2

λ	ωCR	Required β
∞	1·0	10·66
1000	1·0	10·66
100	0·997	10·8
10	0·968	12·4
2	0·874	52·8

These figures show the necessity for keeping the mesh capacitances at least ten times greater than the transistor capacitance.

Acknowledgments

Acknowledgments are made to Mr. W. Makinson (Managing Director) and Mr. G. Hellings (Head of Research) for permission to publish this article and to Mr. N. Wells who helped with the experiments.

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Use of a Transistor for Setting a Square Loop Magnetic Core

By I. Krajewski*, B.Sc.

After showing that most available transistors have to be regarded as voltage sources when setting cores, an analysis of the basic circuit for setting a core is given; this shows that, under given conditions, for fastest setting there is an optimum number of turns for the setting winding. An experimental verification of these results is given which shows quite good agreement with calculation. A practical circuit is briefly described together with calculations of the junction power dissipations within the transistor used.

(Voir page 527 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 532)

WHEN setting a square loop magnetic core it is usual, but not essential, to have a current source. This is because during setting, the core develops back e.m.f. which opposes that trying to set it. To achieve a current source the voltage used for setting the core should be much greater than the back e.m.f. developed by the core being set.

Most available fast switching transistors have relatively low voltage and current ratings, and therefore in general must be considered as voltage and not current sources when setting square loop magnetic cores. It is the purpose of this article to show that there is an optimum number of turns which gives the fastest setting for a given voltage, current limiting resistor and core.

The circuit to be analysed is shown in Fig. 1.

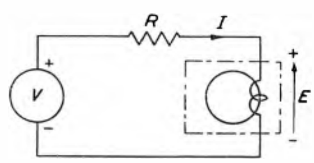


Fig. 1. The basic circuit for setting a core from a voltage source

The dotted line Fig. 1 encloses a square loop core which has N turns wound on it. The e.m.f. E developed across the core due to the change of flux in it during switching is given by:

$$E = -N\,(d\phi/dt) \qquad (1)$$

according to Faraday's Law. This may be approximated to the relation:

$$E = -(N\phi/t) \qquad (2)$$

where ϕ is the total change of flux in webers, and E the average value of e.m.f. developed over the switching time t; E and t being expressed in volts and seconds respectively.

The circuit equation for Fig. 1 is:

$$E + IR - V = 0 \qquad (3)$$

whence:

$$I = \frac{V - E}{R}$$

Substituting the value of E from equation (2):

$$I = \frac{Vt - N\phi}{Rt} \qquad (4)$$

Now the relationship between the switching time t and the driving ampere-turns NI can be expressed approximately in the form.

$$1/t = \frac{NI - A}{B} \qquad (5)$$

which may also be familiar in the form:

$$(H - H_c)\,t = S \qquad (6)$$

and $A = 10\,LH_c/4\pi$ and $B = 10\,LS/4\pi$: L, S and H_c being the magnetic path length, the switching constant and the coercive force respectively.

Substituting for I from equation (4) into equation (5) and re-arranging:

$$1/t = \frac{NVt - N\phi - RAt}{BRt} \qquad (7)$$

cross multiplying:

$$BRt = NVt^2 - N^2\phi t - RAt^2$$

$$\therefore\ BR = NVt - N^2\phi - RAt$$

$$\therefore\ t = \frac{BR + N^2\phi}{(NV - RA)} \qquad (8)$$

It will be evident from equation (8) that there exists an optimum number of turns N for fastest switching. Differentiating equation (8) with respect to N:

$$(dt/dN)\,NV + tV - (dt/dN)\,RA = 2N\phi$$

$$\therefore\ (dt/dN)\,(NV - RA) = 2N\phi - tV$$

$$\therefore\ (dt/dN) = \frac{2N\phi - tV}{(NV - RA)}$$

Substituting for t from equation (8) and rearranging:

$$(dt/dN) = \frac{2N\phi\,(NV - RA) - (BRV + N^2\phi V)}{(NV - RA)^2}$$

Setting $(dt/dN) = 0$ for a minimum then:

$$2N^2\phi - 2N\phi RA - BRV - N^2\phi V = 0$$

$$\therefore\ N^2\phi V - 2N\phi RA - BRV = 0$$

$$\therefore\ N^2 - (2RA/V)\,N - (BR/\phi) = 0$$

Solving this equation for N:

$$N = 1/2\left[2RA/V \pm \sqrt{\left(\frac{4R^2A^2}{V^2} + \frac{4BR}{\phi}\right)}\right]$$

The negative sign is inadmissible since it will make N negative.

$$\therefore\ N_{(opt)} = (RA/V) + \sqrt{\left(\frac{R^2A^2}{V^2} + RB/\phi\right)} \qquad (9)$$

Experimental verification of equations (8) and (9) was carried out on a Plessey 2·4mm/S4 magnetic core, using a Mullard OC42 transistor as the driver. The following constants as quoted by the makers were used in equation (8).

$A = 0{\cdot}4$

$B = 0{\cdot}63 \times 10^{-6}$

$\phi = 94 \times 10^{-9}$ Webers

Fig. 2 shows a graph of t plotted against N using equation (8), as well as an experimental curve, for

$V = 16\text{V}$

$R = 150\Omega$

* *International Computers and Tabulators Ltd.*

It will be observed that the agreement between theoretical and experimental results is quite good, bearing in mind the approximations involved. Both graphs indicate a broad minimum, giving considerable margin in the choice of the number of turns.

Circuit Description

Fig. 3 shows the circuit used in the above investigations. The circuit is a straightforward amplifier, and it is not proposed to give in detail the operation of the circuit. The

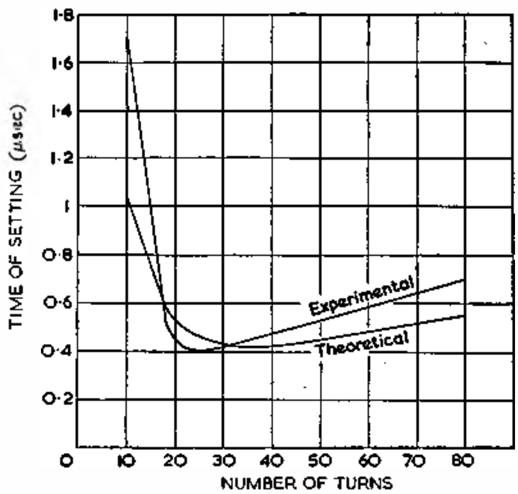


Fig. 2. A comparison of plots of switching time against number of turns on the setting winding

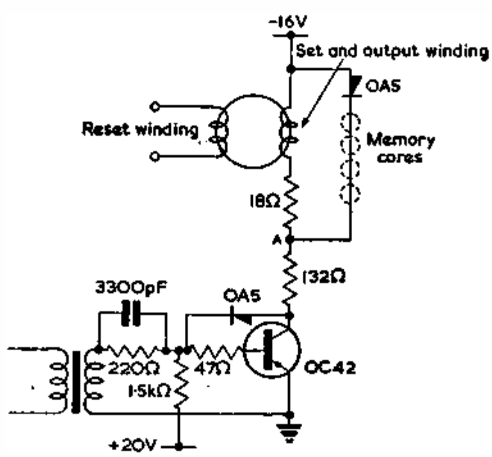


Fig. 3. A practical circuit

points of particular interest about the circuit are the following:

(1) The use of a diode between collector and base to prevent excessive hole storage.

(2) The use of a diode to prevent excessive voltage being developed at the collector when the core is switched out.

Consider the circuit shown by Fig 4, where all the windings on the core are separate. During resetting, the voltage induced at the collector will be of such a polarity as to bring it more negative than its quiescent value of −16V. This is highly undesirable and to prevent this happening the 'set' and 'output' windings are combined, as shown in Fig. 3 so that during resetting any negative excursion of point (A) is suppressed by the OA5 diode which also serves to route the resetting current into the memory cores.

Dissipation

It is of interest to determine the maximum p.r.f. of which the circuit is capable. This is dependent on the frequency cut-off and the dissipation of the transistor.

An estimate of the dissipation of the transistor, working in the circuit shown in Fig. 4 is given below.

The voltage and current waveform of the collector are approximately as shown in Fig. 5.

The average power per cycle P is:

$$P = \frac{1}{t_1 + t_2 + t_3}\left[\int_0^{t_1} (I_c t/t_1)\big(V_o - (V_c t/t_1)\big)\, dt + \int_0^{t_2} \big(I_o - (I_c t/t_2)\big)(V_c t/t_2)\, dt + \int_0^{t_3} V_{ce} I_c\, dt\right] \ldots (10)$$

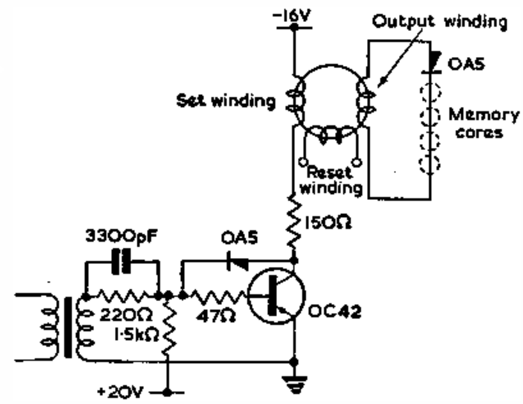
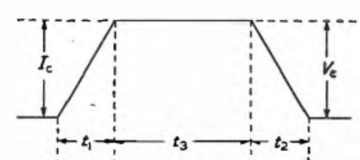


Fig. 4. The arrangement in Fig. 3 modified to prevent an excessive excursion of collector potential

Fig. 5. Symbols denoting intervals of a pulse



Rearranging equation (10):

$$P = \frac{1}{t_1 + t_2 + t_3}\left[V_c I_c \int_0^{t_1} (t/t_1^2)(t - t_1)\, dt + V_c I_c \int_0^{t_2} (t/t^2)(t - t_2)\, dt + V_{ce} I_o \int_0^{t_3} dt\right]$$

Integrating one obtains:

$$P = \frac{1}{t_1 + t_2 + t_3}\left[\frac{V_c I_o}{6}\, t_1 + \frac{V_c I_o}{6}\, t_2 + V_c I_c\, t_3\right] \ldots\ldots\ldots\ldots (11)$$

In the case of the circuit shown in Fig. (4):

$V_o = 16\text{V}$

$I_o = 100 \times 100^{-3}\text{A}$

$V_{ce} = 0{\cdot}28\text{V}$

$t_1 = 0{\cdot}2 \times 10^{-6}\text{sec}$

$t_3 = 0{\cdot}45 \times 10^{-6}\text{sec}$

$t_2 = 0{\cdot}15 \times 10^{-6}\text{sec}$

Substituting those results in equation (11) one finds that the average power dissipated per cycle is 133mW. Working at

500kc/s the average power dissipated would be $\frac{133 \times 0{\cdot}8}{2}$ $= 53$mW.

Similar calculations for the base junction indicates that it dissipates about 3mW; thus the total dissipation of the transistor would be $53 + 3 = 56$mW. Now, the maximum permissible dissipation of an OC42 at 30° ambient is 65mW, indicating that 500kc/s is almost the highest frequency at which the circuit can be used.

Sometimes it may be required to determine the maximum frequency at which a circuit can be operated, given the rise, fall and duration of the pulse and the maximum average dissipation of the transistor P (max).

Let $t_1 + t_2 + t_3 = \tau$

Let $f = (1/T) =$ the frequency of operation; the ratio of τ/T is the duty cycle, being the ratio of the period of operation to the period of rest of the transistor. Since P is the average dissipation per cycle, the average dissipation of a transistor with a given duty cycle is:

$$\frac{P \cdot \tau}{T} = P_{av} \qquad (12)$$

Now let $P_{av} = P_{max}$

$$\therefore \frac{P \cdot \tau}{T} = P_{max}$$

$$\therefore 1/T = f = \frac{P_{max}}{P \cdot \tau} \qquad (13)$$

Equation (13) then determines the maximum frequency of operation of which the transistor is capable without exceeding its dissipation.

For the circuit under discussion:

$\tau = 0{\cdot}8 \times 10^{-6}$sec

$P = 133$mW

P_{max} at 30°C $= 65$mW

$$\therefore f = \frac{P_{max}}{P \cdot \tau} = \frac{65}{133 \times 0{\cdot}8} \times 10^6 = 0{\cdot}60 \times 10^6\text{c/s}$$

Acknowledgements

The author wishes to thank the Directors of International Computers and Tabulators Ltd for permission to publish this article.

The author also wishes to thank M. L. N. Forrest for helpful discussions and encouragement in preparing this article. Thanks are also due to Guido Wattson for suggesting the use of a diode to suppress the negative excursion of collector voltage during resetting.

REFERENCE

REINER, J. Magnetic Cores for Airborne Digital Computers. *Proc. Nat. Electronics Conf.*, *13*, 653 (1957).

Analogue-Digital Computing System

A combined analogue and digital computing system has been developed by Redifon Ltd of Crawley, Sussex. Known as RADIC (Redifon Analogue-Digital Computing System) this new system has been designed for general purpose use as a system rather than a fixed size computer.

Basically the system comprises an assembly of d.c. drift-corrected amplifiers, electronic multipliers and diode-network function generators, operating on a relatively fast time scale. These electronic analogue computing elements, backed up by electro-mechanical servo-operated integrators and multipliers, perform the operations of addition and subtraction, multiplication and division, integration and differentiation with respect to time and the generation of mathematical functions (e.g. sines, squares logarithms).

The different computing elements can be interconnected in various ways, depending on what type of computation is to be performed, by means of a central patchboard.

For feeding numerical data into and out of the analogue computing elements, and also for storage of data, digital devices are used. There are two keyboards for input information. One, on a modified "Addo X" digital printer, is for feeding numerical data into the storage system. The other keyboard is for the automatic setting of coefficients in the analogue computing elements. There are also two devices for digital output information—an illuminated number display (on both the main computer and the central desk) and a print-out mechanism (on the "Addo X" printer). Digital storage is provided by a magnetic tape equipment, which can also be used for the simulation of time delays and other special purposes.

All these digital input, output and storage devices operate in the parallel mode which enables the digital units to match the high speed of the analogue computing elements. The translation of data from digital to analogue form and vice versa is done by analogue-to-digital and digital-to-analogue convertors.

A feature of the RADIC system is the magnetic-tape digital storage system. This extends the field of use of the system to complex types of input data which the normal type of analogue computer is unable to handle. Physically this storage system takes the form of a separate transportable cabinet which contains two tape decks, a digital encoder for translating input data into a form suitable for storage, and a digital-to-analogue convertor for translating digital data from the tape into a form suitable for the analogue computing elements.

A special version of the unit, designed to operate as a self-contained, transportable process analyser, will be available. This will contain, in addition to the above facilities, multipliers and integrators to allow auto-correlation or cross correlation functions to be computed from analogue data recorded on the tape or tapes from the processes in question.

A new principle is used for the storage of digital information on the magnetic tape. This is the principle of quantizing (separating into small steps) the time scale as well as the data recorded. In a conventional magnetic tape data storage system the time scale is continuous, that is, the tape moves in a steady fashion and the data is recorded and read out in a continuous stream. In the RADIC equipment the magnetic tape, which is sprocketed like cinematograph film, is moved in separate small steps, and digits are recorded and read-out while the movements are taking place.

By this means the stored information is put under more precise control than it would be on a continuously moving tape, and, in fact, the small steps are accurately synchronized to the read in/read out circuits. Although the tape always moves at constant speed over the magnetic heads, the intervals of time between successive movements can be extended at will. As a result the average tape speed can be precisely varied over an extremely wide range of speed. The electrical pulse which controls the stepping action can be set at a constant rate or made to vary in accordance with some other quantity involved in a computation.

One special use of this storage system is that it can be used to simulate time delays which may occur in process control systems. Data is recorded on the tape and then read out by a reading head which is physically displaced from the recording head—the time taken for a point on the tape to pass from the recording head to the reading head being determined by an accurately controlled number of steps, each of which takes an accurately controlled interval of time. In addition this time-delay facility is used for correlation computing, in which data stored on the tape is continuously compared with data on other parts of the tape (displaced by known intervals of time) in order to detect periodicities. For this work it is necessary to be able to automatically alter the spacing between the two heads by an electrical signal to give a series of different time delays as the computation proceeds. This is done by a mechanism in the tape deck, based on differential movement between two tape driving sprockets.

On the control desk, besides the digital input and output equipment, there is also a bank of push-buttons for the complete overall control of the computer.

Short News Items

At the British Scientific Instrument Research Association's headquarters at Chislehurst, Kent, work has now started on the erection of a new building. The building, which will comprise new workshops and mechanical laboratories, should be completed in the late autumn. The new laboratories will enable the Association to extend its activities in the mechanical engineering field, while in the workshops provision has been made to accommodate more specialized equipment and precision machine tools, thereby improving the workshop facilities available both to the laboratories and to member firms in the Association.

At a private ceremony at Marconi House, London, on 6 July, Lord Nelson of Stafford, Chairman of The English Electric Group, presented to the National Trust the title deeds of some 40 acres of land at Poldhu, Cornwall. The gift was accepted by Earl De La Warr, Chairman of the Trust's Estates Committee. It was at Poldhu that Guglielmo Marconi erected his radio station which on 12 December 1901 transmitted signals across the Atlantic which were received by him at St. John's, Newfoundland. The Poldhu station was dismantled in 1937 when a granite column was erected to mark the site.

The Electronic Engineering Association and the Radio Industry Council are again sponsoring the annual awards which have been made for the past ten years for published articles on electronics. Articles published during 1960 by other than professional writers will be considered by a panel of judges now consisting of:

- Air Marshal Sir Raymund G. Hart, Director R.I.C.*
- Professor H. E. M. Barlow, Prof. of Electrical Engineering, University College, London.
- Mr. B. C. Brookes, Senior Lecturer, Department of Engineering, University College, London.
- Mr. A. H. Cooper (E.M.I. Ltd).
- Mr. F. Jeffery (Murphy Radio Ltd).
- Mr. G. Reeves (A.E.I. Ltd).
- Dr. R. C. G. Williams (Philips Electrical Ltd).

Articles published in trade and technical journals serving user industries are eligible as well as those in radio and electronic journals. Six premiums of 25 guineas each are offered to the writers of what are judged to be the best articles published between January and December this year. Further particulars can be obtained from the Secretary, The Electronic Engineering Association, 11 Green Street, Mayfair, London W.1.

* *The death was announced on 17 July of Sir Raymund Hart.—Ed.*

The BBC's Television Centre at the White City, which officially opened on 29 June, is believed to be the biggest television headquarters in Europe, occupying a site area of some 13 acres. Its technical and production resources are second to none. Studio Three, which is the first to be commissioned, is a six-camera studio. For this the BBC is using Marconi Mk IV image orthicon camera channels equipped with the English Electric Valve Company's 4½in camera pick-up tubes. Marconi have, in addition, supplied the following items for the Television Centre:

- 13 Mark IV image orthicon television camera channels.
- 11 Mark IV vidicon camera channels.
- 75 21in monitors for use in the production areas.
- 10 Slide projectors, complete with remote control panels and random selection remote control panels.
- 6 Optical multiplexers.
- 5 Electronic switches, together with pattern generators.
- 3 Picture and waveform monitors.

An extensive network of telecommunication cables has been installed by Standard Telephones and Cables Limited for the British Post Office to interconnect the new Television Centre with other BBC centres, at Broadcasting House and a number of buildings in the Shepherd's Bush area, and GPO switching points at Shepherd's Bush, Acorn, Faraday and Museum exchanges. This has, therefore, involved the installation of many new cables for telephone, broadcast sound and video transmission. Among the many types of cable supplied for television picture transmission are 2-, 4- and 6-core coaxial cables with a tube diameter of 0·375in, 3-core coaxial cable with a tube diameter of 0·975in, video pair cable and composite audio pair/coaxial cables.

Steatite and Porcelain Products Ltd, Stourport-on-Severn, have announced that an agreement has been concluded with the Compagnie Generale d'Electro-Ceramique for full technical collaboration, including exchange of patent rights. At the same time, Steatite and Porcelain Products have become selling agents in Great Britain for the products of Compagnie Generale d'Electro-Ceramique, which company will now represent Steatite and Porcelain Products in France. In this way, special ceramics and insulator designs from the French company will be made immediately available to the British market, until manufacture is established at the Stourport factory of Steatite and Porcelain Products Ltd.

The Simms Motor and Electronics Corporation have announced that they have acquired the whole of the issued share capital of Dawe Instruments Ltd of Ealing, Middlesex, and their associated manufacturing company, LMK Manufacturing Co. Ltd of Brentford. These developments expand still further the electronics side of the Simms Group which only recently was augmented by the acquisition of Messrs. Cawkell Research & Electronics Ltd. With the Cawkell and Dawe interests now under the same control, the Simms Group is in a position to provide complete coverage on component manufacture, applied electronics and instrumentation.

The fifth Reactor School course on the Control and Instrumentation of Reactors will be held from 12 to 23 September 1960, and will be open to British and overseas students. It will take place at Durley Hall, Bournemouth, Hampshire. The course is primarily intended for those who have a direct interest in the control and instrumentation of nuclear reactors, and it will be assumed that participants have some knowledge of the basic principles of these subjects. In addition to lectures there will be visits to zero energy and high flux research reactors and other relevant parts of the Harwell Establishment. There will also be a visit to the Atomic Energy Establishment, Winfrith.

Associated-Rediffusion's new Studio Five, which has recently been opened, is the largest television studio in Europe and the British Commonwealth. EMI Electronics Ltd have supplied all the television camera channels and associated video equipment used in the studio, and have also been responsible for the technical wiring and installation of all technical equipment. The company have supplied 10 of the new EMI image orthicon camera channels, 50 21in picture monitors, camera control and monitoring equipment, and duplicated vision mixing. A special feature of the studio is a soundproof partition which can be raised or lowered to divide the studio into two. Each half of the studio is equipped with its own vision production and control facilities to provide facilities for the operation of either side separately, or for the integration of both halves into one overall operation. The Studio can function on 405, 525 or 625 line standards and the vision equipment is designed to operate on any one of the three standards by the use of a switch or changeover plug connector, without the need for supplying any substitute units or carrying out any modifications. The Studio can be converted from one standard to another in less than 20 minutes.

Investigations into the behaviour of materials at very low temperatures are being carried out at the Clarendon Laboratory of the University of Oxford. Of special importance is the variation in the magnetic properties of materials at temperatures approaching absolute zero. To extend the University's efforts in this direction a new laboratory is to be built for the purpose and Mullard Ltd have made a grant of £25 000 towards the cost of the new building. The new laboratory will be known as the Mullard Cryomagnetics Laboratory.

The first direct telephone cable between the United Kingdom and Sweden is now being installed between Middlesbrough and Gothenburg. It is the longest submarine telephone cable system in Europe and is also, to date, the longest in the world using a single cable for both directions of transmission, being some 530 nautical miles in length. The polythene dielectric coaxial cable was manufactured at the Standard Telephones and Cables works at Southampton and was fed direct on to cable ships *Monarch* and *Ariel*. Twenty-nine submerged repeaters installed at intervals along the cable were manufactured at the STC North Woolwich factory in special air-conditioned areas designed especially for the purpose. The repeaters used on the new cable are of the same type as those which will be supplied for CANTAT, the 1961 UK-Canada cable, the first link in the proposed round-the-world Commonwealth cable. Like the CANTAT system repeaters, they are designed to carry 60 telephone circuits of 4kc/s spacing. The system is expected to be in service in October of this year.

The Council of the Royal Society has appointed Dr. D. T. Edmonds, of the Clarendon Laboratory, University of Oxford, to a Mackinnon Research Studentship from 1 October 1960 to enable him to carry out magnetic investigation below 1°K in large magnetic fields at the Clarendon Laboratory, Oxford.

A new Society called 'The Biological Engineering Society' was inaugurated at a recent meeting held at the Royal College of Surgeons. The meeting was attended by doctors, physiologists, electronic engineers, mechanical engineers and physicists from many parts of the country. The purpose of the Society is to bring together members of these different disciplines, from hospitals, research institutes and industry, to further the applications of engineering to biological and medical problems. Dr. R. Woolmer, Professor of Anaesthesia at the Royal College of Surgeons, was elected first President. The Acting Secretary is Dr. A. Nightingale, Physics Laboratory, St. Thomas's Hospital, London, S.E.1, who will be glad to receive inquiries about the new Society.

Over 2 000 members and guests attended a Conversazione held at the Royal Festival Hall on 23 June, and were received by Sir Willis Jackson, F.R.S., president, and Lady Jackson, assisted by Sir Hamish D. MacLaren, K.B.E., C.B., D.F.C., president-elect, and Lady MacLaren and members of the Council of The Institution of Electrical Engineers.

The incorporation of the Institute of Physics and The Physical Society into a new body to be known as 'The Institute of Physics and the Physical Society' has been announced. The first President of the amalgamated society is Sir John Cockcroft and the executive officers are: Secretary, Dr. H. R. Lang; Editor and Deputy Secretary, Dr. A. C. Stickland; Deputy Secretary, Mr. N. Clarke. The registered offices and headquarters are at 47 Belgrave Square, London, S.W.1, and for the present, offices will also be retained at 1 Lowther Gardens, South Kensington, London, S.W.7.

Wing Commander A. R. Gilding has been appointed Assistant Technical Secretary of the Electronic Engineering Association, joining from the Air Ministry, where latterly he was in charge of the branch responsible for airborne radio servicing policy and some aspects of new airborne equipments. Wing Commander Gilding specialized in communications and radar throughout his career in the R.A.F., which began in 1927.

The Telegraph Condenser Company Ltd of the United Kingdom, and Sprague Electric Company of the U.S.A. announce that they have come to an arrangement whereby TCC acquire the right to all Sprague United Kingdom patents and applications, together with the technical and engineering information necessary to exploit them. Sprague will make available to TCC all its present research and technical information and engineering knowledge and for the next 21 years the two companies will exchange research, development and manufacturing knowledge, extending beyond the field of capacitors and embracing all the products of the two companies.

The death was announced of Air Marshal Sir Raymund Hart, K.B.E., C.B., M.C., on 17 July at his home at Aston Rowant, Oxfordshire. Sir Raymund Hart, who was 61, was Controller of Engineering and Equipment at the Air Ministry until his appointment as Director of the Radio Industry Council in March 1959. He was one of the early pioneers of radar and during the war was responsible for the development of the operational use of radar on the ground and in the air. He was Chief Air Signals Officer, Allied Expeditionary Air Force, during the planning and operational stages of the allied invasion of Europe in 1944.

PUBLICATIONS RECEIVED

BRITISH INSTRUMENTS, Second Edition, is published by the United Science Press Limited under the authority of the Scientific Instrument Manufacturers' Association, SIMA House, 20 Queen Anne Street, London W.1. This edition has expanded from a total of 615 to 635 pages. The foreign glossaries in French, German and Spanish to assist overseas users appear again, and also the classified list of instruments and components, list of trade names, manufacturers' reference sheets, description of the associations allied to the instrument industry, and an alphabetical list of instrument and component manufacturers with overseas agents. Price 45s. post free.

THE STORY OF A FAMILY FIRM, by M. Bache, gives the history of the firm of Geo. Salter & Co. Ltd of West Bromwich, England, since it was founded in 1760. Price 21s.

BASICS OF MISSILE GUIDANCE AND SPACE TECHNIQUES, by M. Hobbs, is Volume I of a series of books written to provide a comprehensive basic course in missile guidance and related space systems in fundamental terms. They are directed to the educational needs created by the rapidly growing requirement for technical and semi-technical personnel to satisfy present and future missile and space vehicle programs, as well as to those wishing to learn more about this subject on an informal basis. John F. Rider Publisher, Inc, New York. Chapman & Hall Ltd, London. Price 63s.

RECIPROCALS OF THE INTEGERS FROM 1000 to 9999, by T. H. Redding, is a book containing tables listing the 10^8 multiples of these reciprocals; they are direct reading and the entries are correct to the nearest integer in the third place of decimals. Although suitable for general use, the tables have been prepared and arranged so that they are particularly suitable for use in conjunction with mechanical computing machines of the 'barrel-setting' type in the evaluation of quotients and compound fractions. Equally, the tables find application in effecting the evaluation of quotients on adding and listing machines when these incorporate a mechanism for automatically evaluating simple products. There is a 21-page Appendix which prefaces a discussion of the tables with a description of barrel-setting machines and their simplest method of operation in the evaluation of products, quotients and compound fractions. Taylor & Francis Limited, Red Lion Court, Fleet Street, London E.C.4. Price 18s. 6d.

I.C.T. 1301 DATA PROCESSING SYSTEM is a brochure issued by the Publicity Division, International Computers and Tabulators Limited, Gloucester House, 149 Park Lane, London W.1. This brochure has been divided up into three sections to cover different approaches to the system namely: (a) A layman's guide to the computer, for the non-technical business executive. (b) A semi-technical description of the computer's principal units and features, with details of the I.C.T. services available to all who are interested in its use. (c) Complete detailed specification and order code, providing the sort of information demanded by the technical reader.

EARPHONE INSETS FOR GENERAL USE is an illustrated technical leaflet describing miniature magnetic earphones for use in hearing aids, listening units, lightweight headphones, telephone handsets and radiation detectors. Amplivox Limited, Industrial Division, Beresford Avenue, Wembley, Middx.

PRECIOUS METAL CONTACTS is the title of a new catalogue issued by Engelhard Industries Ltd, Baker Platinum Division, 52 High Holborn, London W.C.1, which should be useful to designers and engineers in the electrical industry in selecting the most suitable precious metal alloy from the many which are currently available. Details are given of the standard types of contacts and fabricated parts used in the electrical industry, together with technical data and properties of many precious metal alloys currently available. Physicists, metallurgists, chemists and engineers of the Engelhard Organization are available for consultation on any problem connected with the application of the precious metals which include those of the platinum group as well as of gold and silver.

LETTERS TO THE EDITOR

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

Common-Aerial Transmission and Reception at V.H.F.

DEAR SIR,—In the course of a design study for an h.f. ionosphere sounder, an application has occurred to us which has a close bearing on the topic of Mr. J. K. Grierson's article on a common-aerial system for transmission and reception at v.h.f. (ELECTRONIC ENGINEERING, September 1959). We propose to employ a transmission-line equivalent of the ring-hybrid duplexer[1,2], used in many microwave equipments and the layout is shown in Fig. 1. For operation around 25Mc/s, 600Ω open wire feeders will be used with an 850Ω ring.

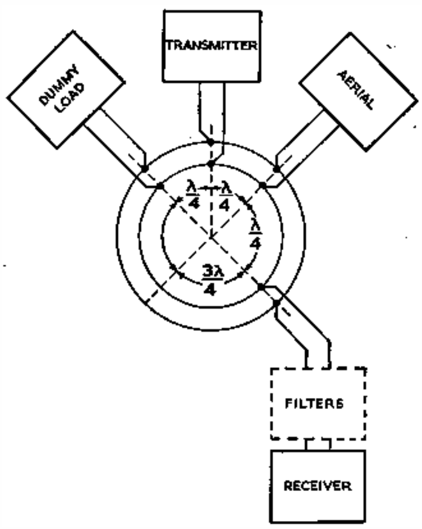


Fig. 1. Transmission line equivalent of the ring-hybrid duplexer.

When the ring is correctly balanced, power from the transmitter divides equally between the aerial and the dummy load; similarly, incoming power divides equally between receiver and transmitter. Ideally, no outgoing power should find its way to the receiver, but in practice, difficulties of balancing make it desirable to insert filters in the receiver path to improve isolation at the transmitted frequency. These may, however, be relatively simple in design, with quite low power-handling capacity.

The obvious advantages of the arrangement are:

(1) The simplicity and cheapness of construction of the outdoor installation.

(2) The lack of sophistication in filter requirements.

Corresponding disadvantages are:

(1) The device is of a narrow-band type, though it would be more than adequate to deal with Mr. Grierson's choice of a 2·5 per cent separation between the transmitted and received frequencies;

(2) half the outgoing power is dissipated in the dummy load, as compared with Mr. Grierson's filter losses of 320 watts out of one kilowatt;

(3) there is an apparent 3dB loss in received power as well.

This third point is, however, less serious in some applications than would appear at first sight. The noise input to the receiver arises partly from the aerial and partly from the dummy load. If at v.h.f., the aerial noise temperature is very high (due to atmospheric noise), then the extra contribution from the dummy load is not serious and can be neglected. Further, the aerial noise power, as well as the signal power, divides equally between the transmitter and receiver paths, so that the effective signal-to-noise ratio will not deteriorate by as much as 3dB. This would definitely be so, for instance, in a communication link employing forward-scatter or allied techniques. The question of whether a loss of 3dB signal power significantly affects the operation of the receiver is quite a different matter which depends primarily on receiver noise factor.

Yours faithfully,

A. J. HYAMS AND J. LAIT,

Royal Military College of Science,
Shrivenham, Wilts.

REFERENCES

1. TYRRELL, W. A. Hybrid Circuits for Microwaves. *Proc. Inst. Radio Engrs. 35*, 1294 (1947).
2. SOUTHWORTH, C. G. Waveguide Transmission. pp. 335-339, (Van Nostrand, New York, 1950).

Mains Frequency and Oscilloscope Time Markers

DEAR SIR,—The accuracy of electric clocks based on the 50c/s mains frequency, when averaged over hours or days, is apt to give a false impression of the accuracy of the same mains frequency when used to measure time intervals of some small fraction of a second. We have been engaged for some years, in this department, on acoustical research using tape-recorded material which is then photographed, using moving film cameras and oscilloscopes. When devices, similar to those described recently[1] were used to give precise time marking signals at intervals of 0·01sec or 0·001sec, variations in the apparent frequencies of the *same* original recorded signals were noted in different oscillograms, although the speed of transport of the photographic film was constant throughout each photograph.

The same effect was found when the time signal was obtained from a thermostatically controlled tuning fork, giving 1kc/s at 25·9°C. Of course we have here the additional complication that the frequency of the tape-recorded signal is dependent also upon the tape speed on replay, but this, in turn, depends upon the mains frequency. The order of magnitude of the effect during the last three years is that the frequency of the same tape-recorded note taken originally from the Royal Festival Hall organ varied by about three vibrations at a mean frequency of 262c/s. Normally the Central Electricity Generating Board is able to keep the mains frequency constant at ±0·2 per cent but on rare occasions during extremely rapid changes of load or during breakdown, the frequency may be as much as 2 per cent below normal, but the System Control Office has records of these changes.

There is little danger in using the mains frequency as a time-marker to check the constancy of speed of movement of the photographic material in an oscilloscope camera or to make relative time measurements on the resulting oscillogram. Errors may occur, however, in absolute measurement of time intervals on the same oscillogram. The anomaly would not have been noticed but for the extreme sensitivity of the ear to small changes of pitch. Some time ago, in preparing a composite recording for a broadcast talk, I found that each of eight disk recordings of the same composition, made on different occasions, when replayed using the same synchronous motor, were recorded in slightly different keys. One older disk replayed a semitone below the correct (A = 440c/s) key. I understand that during important recordings the mains frequency is now always monitored, so that when the recording is made permanent, by being transcribed from tape to disk, any error of timing can be corrected. Caution is therefore always necessary in using the mains frequency for time marking when high accuracy is needed.

Yours faithfully,

W. H. GEORGE,

Department of Physics,
Chelsea College of Science and Technology,
London, S.W.3.

REFERENCE

1. THOMAS, J. J., DAFT, C. H. J. A Simple Time Marker. (Letter) *Electronic Engng. 32*, 117 (1960).

Unusual Characteristics of Silicon Transistors

DEAR SIR,—Some special samples of silicon transistors made by the Philco etching and alloying process (supplied by Semiconductors Ltd) have been found to show a remarkably high current gain at very low collector currents. The following figures were obtained from one specimen:-

Base Current ($\times 10^{-9}$A)	*Collector Current* (μA)	*Base Current* ($\times 10^{-9}$A)	*Collector Current* (μA)
+1·09	0·0	1·01	1·9
1·08	0·2	0·99	2·3
1·07	0·3	0·97	2·8
1·05	0·7	0·95	3·1
1·03	1·3	0·93	3·5

These figures are taken from a series of measurements made on a transistor with a collector-base leakage of about 2×10^{-9}A. For all readings a collector voltage of -3V was used. Averaged over the collector current range 0·7 to 3·1μA the current gain is over 20 000. At higher collector currents the gain falls, reaching a normal value at I_c of 100μA. Out of 60 samples, originally supplied for entirely different properties, 10 showed the same type of characteristics.

One manifestation of the effect is a high ratio of I_{co}' to I_{co}. This was the property first observed and led to the closer examination of the transistors. Similar transistors may be quickly separated from normal types using this as a test.

It appears that the effect is due to there being a very small area within the transistor and showing these high gain properties. This could take place during either the etching or alloying process and be caused by defects in the original silicon crystal or the surface of the wafer. It is not yet known whether these properties can be obtained intentionally. If such devices could be produced commercially they would be particularly useful as input stages for direct-coupled amplifiers. Due to the low current level of operation, current drift would be low and input impedance high.

Yours faithfully,

R. B. D. Knight,
Dorking, Surrey

Analogue Computer Techniques

Dear Sir,—In his article in your February issue (page 74), Mr. F. C. Harbert states that to determine whether the coefficient matrix of a system of linear equations is positive definite or not, 'requires as much labour as solving the system'.

It may be of interest to recall that there exist sufficient conditions which may be tested by inspection. It was shown by Parodi[1] that the eigenvalues of an arbitrary matrix A are necessarily confined to be in one at least of the set of n circles

$$|a_{ii} - p| \leqslant \sum_{j=1} |a_{ij}| \qquad j = 1, 2, \ldots m; \; j \neq i \quad (1)$$

and also in another set

$$|a_{jj} - p| \leqslant \sum_{i=1} |a_{ji}| \qquad i = 1, 2, \ldots m; \; i \neq j \quad (2)$$

p denoting here the complex variable.

For the case considered in the article, for which clearly all coefficients must be real,

$$a_{ii} \geqslant \sum_{i} |a_{ij}| \qquad i = 1, 2, \ldots m; \; i \neq j \quad (3)$$

or alternatively

$$a_{ii} \geqslant \sum_{j} |a_{ij}| \qquad j = 1, 2, \ldots m; \; j \neq i \quad (4)$$

assure the positive definitiveness of A.

Where a system matrix satisfies neither of these conditions, it may be worthwhile to attempt to bring the matrix to a form satisfying them by permutation of rows and columns and if necessary change of sign of some rows or columns.

Mr. Harbert states further that the $4n$ amplifier system will 'always yield stable solutions on an analogue computer'.

While this may be true in most cases it has not yet been generally proved when the physical characteristics of the amplifiers are taken into account. It is only true that the matrix of the enlarged equivalent system is positive definite, but the conditions for stability must be further examined.

The treatment will be limited to positive coefficients, or, alternatively, the phase invertors will be supposed ideal.

Several authors[2,3] have shown that in this case the solutions p_k of the set of equations

$$\frac{n+1}{G(p)} = -\lambda_i \qquad i = 1, 2, \ldots n \quad (5)$$

determine the nature of the transients and hence

$$\mathrm{Re}(p_k) > 0 \quad (6)$$

is necessary to assure stability.

$G(p)$ denotes here the amplification of the amplifier depending on the complex frequency p. The λ_i are the i eigenvalues of the system matrix.

It may be shown[4] that the condition (6) can be expressed as

$$\mathrm{Re}(\lambda_i) > 0 \text{ and } |\lambda_i| < \frac{n+1}{|G_{90}|} \quad (7)$$

or

$$\mathrm{Re}(\lambda_i) > 0 \text{ and } |\lambda| < \frac{n+1}{|G_0|} \quad (8)$$

where G_{90} and G_0 denote the amplification for 90° and 0° phase shift respectively.

For positive definite matrices the first part of condition (7) is satisfied. If one supposes, for convenience, $|a_{ik}| \leqslant 1$ $(i, k = 1, 2 \ldots n)$, the second part of condition (7) making use of condition (3) or (4) takes the form

$$n + 1 \leqslant \frac{n+1}{|G_{90}|} \quad (9)$$

giving

$$|G_{90}| \leqslant 1 \quad (10)$$

Thus under the assumptions made, if the amplifier has a phase margin larger than 90°, the set-up for a system having positive definite coefficient matrices will be stable.

It may be noted that the stability conditions in this form involve no differentiation on the variables and thus the treatment includes cases where the variables are represented by a.c. as well as d.c. voltages.

A. Fuchs,
Scientific Department,
Israel Ministry of Defence.

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2. Raymond, F. H., Sur un type général de machines mathématiques algebriques. *Ann. Telecommunications* 5, No. 1, (1950).
3. Goldberg, E. A., Brown, G. W. An electronic simultaneous equation solver. *J. Appl. Phys.* 19, 339 (1948).
4. Cederbaum, I., Fuchs, A., On the stability of linear algebraic equation solvers. Actes des Secondes Journées Intern. de Calcul. Analogique, (Strassbourg 1958).

The Author Replies:

Dear Sir,—I thank Dr. Fuchs for his comments and information. It is true that the application of analytical tests for positive definiteness is not strictly difficult. The results of tests are, however, difficult to apply in subsequent manipulation of the matrix. Much time and labour can be spent in attempting to bring the matrix to a stable form and in some cases this attempt may prove to be fruitless. It is for this reason that the second method of computer solution is recommended.

My statement that the $4n$ amplifier system will 'always yield stable solutions on an analogue computer' assumes a perfect computer. Instability in a practical system can arise from many causes; inter-coupling between amplifiers, imperfect computer components etc. In general their effect on computer solutions cannot readily be put in a mathematical form and the best a designer can do is to minimize their effects. I am grateful to Dr. Fuchs for drawing my attention to a test which allows one to check on one possible source of instability.

Yours faithfully,

F. Harbert,
Louis Newmark Ltd,
New Addington, Surrey

Regulated Power Supplies

Dear Sir,—Professor Middlebrook's letter (February, 1960) describes a simple and effective way of improving supply performance, but he does not mention a further advantage to be obtained. If a pentode replaces the triode series valve, then the screen may be returned to the

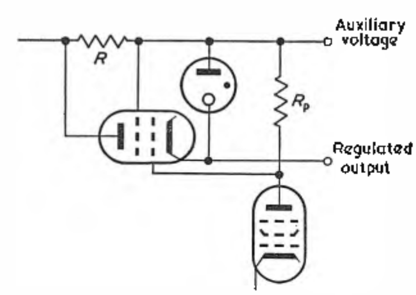


Fig. 1. Use of a pentode valve.

regulated auxiliary voltage (Fig. 1). The high a.c. resistance of the series pentode then gives more attenuation of supply voltage variation than will be obtained with a triode.

Yours faithfully,

Philip Took,
Analytical & Control Instrument Division,
Consolidated Electrodynamics Corporation,
Pasadena, California.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

(Voir page 523 pour la traduction en français: Deutsche Übersetzung Seite 528)

'ON-LINE' ANALOGUE COMPUTER

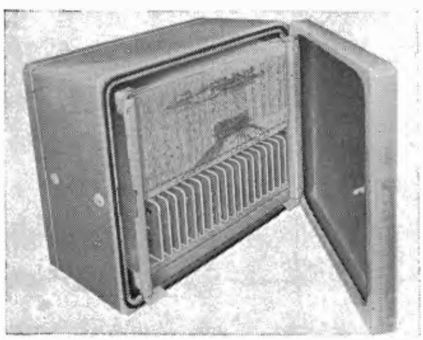
De Havilland Propellers Ltd, Hatfield Aerodrome, Hatfield, Hertfordshire

(Illustrated below)

It is claimed that 'ANATROL' represents a new approach to the design of an on-line analogue computer system and provides for the first time an inexpensive, comprehensive and versatile computer with the stability and robustness vital to modern on-line process control.

The computing accuracy of ANATROL is compatible with present-day measurement accuracies, allowing it to be easily and efficiently integrated into existing control schemes.

Conventional analogue computers provide solutions to process equations at



speeds which are extremely fast when compared with process dynamics. This fact, coupled with the need to increase reliability and to reduce price considerably, led to the choice of the principles used.

The minimum of computing amplifiers and multipliers are formed into an arithmetic unit. This unit is then used to provide a step by step solution of the equations, according to a preset program in which each step is selected sequentially by a switch system. Short stores are used to retain information on the result of a step in the solution, this data being used in a later stage of the computation. Long stores are used to clamp the outputs at the last computed figure while fresh values are being determined.

Up to a maximum of twenty-five steps is available, which gives the computer the potential of a machine employing at least seventy-five computing amplifiers and twenty-five multipliers. Semiconductor techniques are used exclusively, with the exception of some electromechanical switching elements. The machine makes use of module construction, and different component-unit combinations may be used according to the problem. If necessary, an additional input/output unit is available, containing transducers or amplifiers to match the computer accurately to the measuring and control equipment in the plant.

The computer has a built-in self-checking facility and for errors greater than 1 per cent an alarm is given and the computer can be automatically isolated from the process.

EE 24 751 *for further details*

VIBRATION LIMIT CONTROLLER

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

(Illustrated below)

The type 1434 vibration limit controller has been introduced to measure, monitor and record vibration in machines and structures. Large rotating machinery, including both hydro- and turbo-electric plant, as well as heavily loaded bridges and dams can be automatically monitored for excessive vibration by comparing the signal derived from a pick-up continuously with a pre-set bias voltage. A trip valve with a heavy-duty relay in the anode circuit is triggered if the signal rises above the pre-determined level. The relay then shuts down the machine or operates an alarm.

The vibration pick-up is of robust moving-coil design and is hermetically sealed to maintain accuracy under conditions of high temperature and humidity. Twin screened leads up to 200ft long enable the controller to be placed at a safe distance from pick-ups mounted on machines in dangerous or inaccessible locations. Over the frequency range from 20 to 400c/s (1 200 to 24 000rev/min) the pick-up provides an output voltage proportional to the velocity of vibration of the surface with which it is in contact. A signal representing displacement is obtained by applying the output voltage to an integrating network followed by amplification and rectification. The resultant d.c. signal is used to operate a meter calibrated in peak values of displacement for a simple harmonic motion applied to the pick-up. This meter reading of average vibration level can be recorded by plugging in a standard recording d.c. milliammeter.

A sensitivity control provides four ranges: 0 to 0·001in, 0 to 0·003in, 0 to 0·010in and 0 to 0·030in (British units); or 0 to 0·03mm, 0 to 0·1mm, 0 to 0·3mm and 0 to 1·0mm (metric units). The controller can be set to trip at any vibration level between 20 and 100 per cent of the full-scale deflexion of the meter on any range.

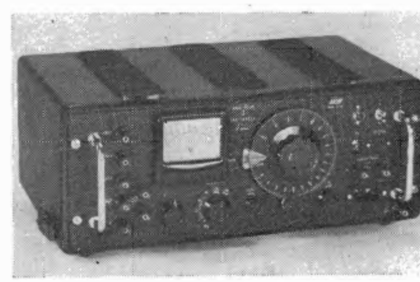
EE 24 752 *for further details*

FREQUENCY ANALYSER

Laboratoire Electro-Acoustique, 5, rue Jules-Parent, Rueil (Seine-et-Oise), France

(Illustrated below)

This instrument has been designed for the analysis of components of a frequency spectrum with a fundamental frequency of 50c/s.



Analysis is carried out as follows:

A variable frequency, produced by a precision generator which is incorporated in the analyser, is superimposed on the spectrum which is to be analysed. After detection the signal is fed to a selective amplifier (with a meter) which only allows a narrow band to pass. Whenever the difference between the variable frequency and the component of a spectrum gives a frequency in the pass-band, a deflexion of the instrument, proportional to the component's amplitude, will be obtained.

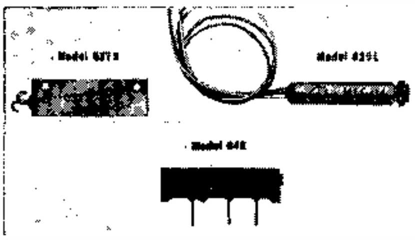
The F.A.H. 17 analyser allows the examination of components, one after another, as well as that of a single component in relation to time.

As the 50c/s fundamental frequency may not be absolutely stable, it is necessary to have a pass-band of a certain width, so that the various components may be tuned. Indeed, should tuning be too sharp, then frequency variations, even very limited, would cause variations in the instrument's reading, without there being, in fact, a variation in the component's voltage.

In addition, it is essential that the instrument's selectivity, apart from the above pass-band, should be sufficiently high to prevent the measurement of a component under examination from being influenced by neighbouring components on either side of the former.

The frequency range is 100c/s to 2kc/s and the attenuation 25dB at ±50c/s from the middle of the pass-band. An output is provided for operating a pen recorder.

EE 24 753 *for further details*



MINIATURE POTENTIOMETERS

Miniature Electronic Components Ltd, The Lye, St. Johns, Woking, Surrey

(Illustrated above)

M.E.C. Ltd are producing under licence from Con-Elco of California a range of miniature trimmer potentiometers available in the three standard configurations illustrated.

The temperature range is −55°C to +180°C, power rating 2¼W, de-rating to ½W at 150°C. The resistance range is 10Ω to 50kΩ. These units have been specially designed and developed for the guided weapons and missile industries, where stringent environmental conditions exist and extreme reliability is all-important.

EE 24 754 *for further details*

A.F. POWER METER

Amos of Exeter Ltd, Weircliffe Court, Exwick, Exeter, Devonshire

(Illustrated below)

The design of the model 157 follows closely upon that of other WEIR-CLIFFE a.f. power meters utilizing a 4½in moving-coil movement scaled 0 to

10mW and −20dB to +10dB. All load resistors, non-inductively wire-wound and rated for continuous full load conditions, are trimmed to an accuracy of 0·1 per cent and provision is made for an independent variable adjustment to compensate for the individual variation in meter movements.

Input impedance matching is by a 'C' core transformer of generous proportions having a ·004in strip core to ensure linear frequency response. A centre-tap is provided so that a fully balanced input is available at all impedances and the insulation is such that the instrument may be directly connected into a valve circuit to determine the optimum load conditions. The two basic impedance controls select values from 1Ω to 10Ω in 0·5Ω steps while a multiplier provides factors of ×1, ×2, ×5, ×10, ×20, ×50, ×100, ×200, ×500 and ×1 000 permitting rapid selection of any impedance from 1Ω to 10kΩ.

A five position switch, directly calibrated, selects ranges of 10mW, 100mW, 1W, 10W and 50W f.s.d., the lowest useful reading on each range being one tenth of full scale.

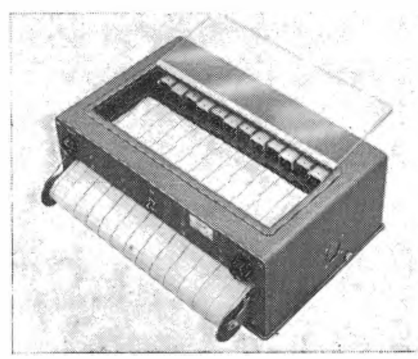
EE 24 755 *for further details*

TWELVE CHANNEL RECORDER

AB Elektrisk Malmletning, Danderydsgatan 11, Stockholm

(Illustrated below)

The ABEM series of direct-writing pen recorders are precision instruments for direct registration of between one to twelve low-speed events; these events may be fed into the centre-tapped



moving coil galvanometers as either a.c. or d.c. signals. The direct recording is obtained by using heated pens and wax covered lined paper. As the paper passes under the pens, the wax surface in contact with them is melted and the bright red paper base becomes visible giving a high contrast trace. Two event markers are provided and may be used to record time or identification marks.

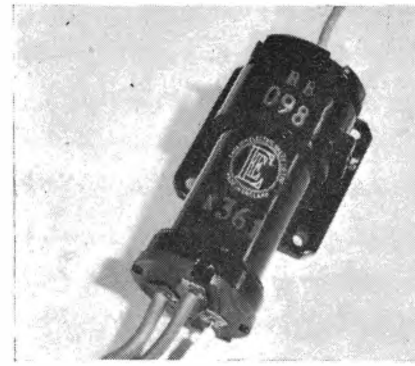
By interchanging gear wheels any three complementary speeds may be obtained within the range of 1mm/min to 10cm/sec. There is a choice of galvanometers with internal resistances of 1 750Ω, 430Ω, 100Ω, 28Ω with sensitivities of 25mm/1mA, 25mm/2mA, 25mm/4mA and 25mm/8mA respectively. The recording paper is in rolls 90m long and is 383mm wide. Pens have a maximum deflexion of 25mm on either side of a central zero reference.

EE 24 756 *for further details*

REFLEX KLYSTRON OSCILLATOR

English Electric Valve Co. Ltd, Chelmsford, Essex

(Illustrated above right)



The K361 low voltage reflex klystron is designed specifically for use in doppler speed measuring systems where reliable performance with minimum routine attention are primary requirements.

Typical applications include an equipment produced by the Westinghouse Brake and Signal Company for automatically controlling the speed of loose-shunted rolling stock in railway marshalling yards, and in Police 'radar' doppler speed measuring systems. Mechanical tuning of this klystron is effected, over the frequency range of 10 700 to 10 725 Mc/s, by a single screw tuner intruding into the cavity.

The K361 has a minimum output power of 20mW and is fitted with a coupler type UG-40 A/U (Z830051) for connecting to No. 16 waveguide.

The heater power required is 6·3V at 0·6A and typical operating conditions include a resonator voltage of 300V, resonator current of 25mA and a reflector voltage of −200V.

EE 24 757 *for further details*

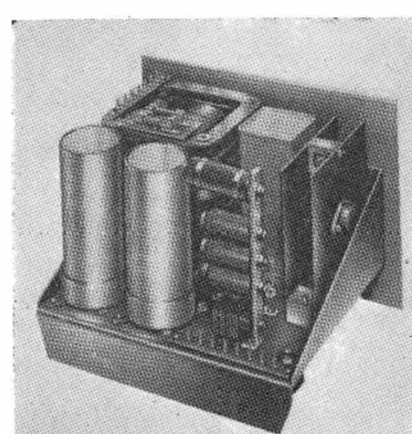
STABILIZED POWER SUPPLIES

International Electronics Ltd, 132-135 Sloane Street, London S.W.1

(Illustrated below)

These units are fully transistorized and are built on a 16 s.w.g. zinc plated steel chassis and provision is made for mounting to a panel or base plate. Any two units can be mounted side by side on a standard 19in rack panel.

The units in the present production range cover currents of 100mA, 200mA, 350mA and 500mV at standard voltages of 200V, 250V and 300V. In addition, all units have two 6·3V 5A heater windings, one of which is centre-tapped.



Full protection of the units is afforded by mains and output fuses.

The output impedance at 500kc/s is less than 0·5Ω, the r.m.s. ripple less than 1mV and the stability factor 400:1.

The first of a range of transistorized low voltage power supply units is also available. This unit will provide a current of 2A at any voltage between 5V and 50V and employs a switching technique to avoid the use of a large number of transistors in parallel. Units providing 5A and 10A will be added to the range in the near future.

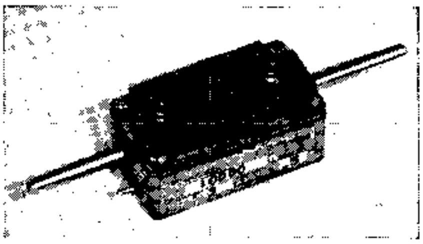
EE 24 758 *for further details*

RECTILINEAR POTENTIOMETER

General Controls Ltd, 13-15 Bowlers Croft, Honywood Road, Basildon, Essex

(Illustrated below)

This precision rectilinear potentiometer with balanced suspension wipers has been introduced to overcome the problem of wiper separation from the resistance element, under extreme environmental conditions of shock and vibration. The unit has a 'U' shaped



wiper member which is held in a pressure equilibrium between the commutator rod and resistive element. Any external forces which tend to alter the wiper contact pressures against the resistive element causes an equal and opposite reaction at the commutator side of the 'U' shaped wiper, thereby ensuring that constant wiper arm pressure is maintained. Such a system holds contact noise and pick-off signal discontinuity to a minimum and prevents damaging wiper pressure against the resistive element.

The maximum number of tappings, including end connexions, is five and the resistance range 100Ω to 10kΩ.

EE 24 759 *for further details*

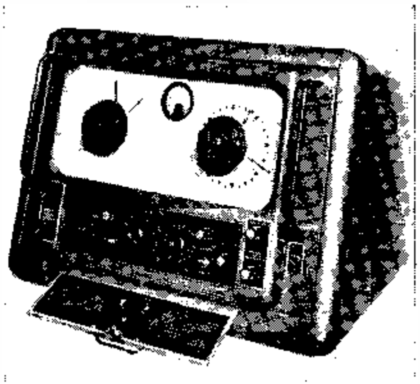
AUTOMATIC M.F. DIRECTION

Marconi's Wireless Telegraph Co. Ltd, Chelmsford, Essex

(Illustrated above right)

The Type ND.103 direction finder embodies the Bellini-Tosi fixed-loops system, and the rotatable element is an automatically adjusted goniometer coil, operating on the null-signal method.

A sensitive receiver and radio goniometer are housed in a single compact unit, suitable for bench mounting. Although designed primarily to function automatically, manual controls are also fitted, under a hinged plate, the essential controls for automatic use of the unit being readily accessible and conveniently situated.



The front panel, sloped for easy reading, carries to the left the tuning scale, and to the right the bearing indication scale, while between the scales is the tuning meter.

The equipment is suitable for fitting with all leading makes of gyroscopic repeaters, thereby obviating the necessity of plotting ship's head, and has been designed to comply with British General Post Office specifications for direction finding equipment for ships in all respects.

The accuracy of bearings on 'automatic' is better than ±1 per cent and an internal crystal calibrator is incorporated for scale checking.

Provision is made for immediate switching from 'auto' to 'manual', enabling bearings to be taken of weak signals in the presence of stronger signals, a situation that might arise when dealing with a distress call.

The output of the receiver is taken to a loudspeaker, but headphone jacks are also provided.

EE 24 760 *for further details*

HIGH TEMPERATURE P.V.C. SLEEVING

Permanoid Ltd, New Islington, Manchester, 4

For high temperature applications or processing Permanoid have introduced a new p.v.c. sleeving with viscose silk internal reinforcement; it is manufactured to Type 1, Class 105T, B.S.2848 : 1957.

Permanoid v.s. sleeving combines the advantages of an integral textile core and p.v.c. sheath without any of the disadvantages. Viscose silk internal reinforcement affords flexibility in use and eliminates shrinkage risk in baking. A plasticized p.v.c. outer coating provides increased density resulting in improved resistance to impregnation oils or varnishes effectively reducing the possibility of mechanical breakdown during processing. It has good ageing properties, is impervious to tropical mould, etc., and is suitable for continuous operating temperatures up to 105°C.

It is available in eleven colours (including black and white) in bore diameters from 0·5mm to 5·0mm.

When tested in accordance with B.S. 2848, the insulation resistance is not less than 1 000MΩ and the breakdown voltage 3kV.

EE 24 761 *for further details*

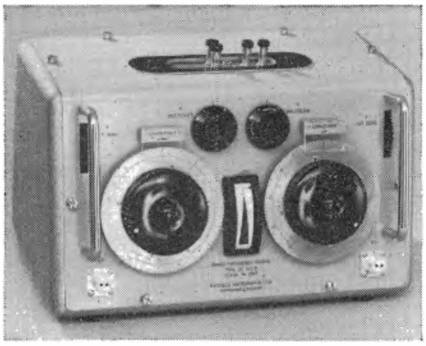
R.F. BRIDGE

Hatfield Instruments Ltd, Crawley Road, Horsham, Sussex

(Illustrated below)

The LE300 r.f. bridge provides, in five ranges, direct calibration of resistance from 10Ω to 10MΩ, capacitance from 0 to 0·025μF (lowest range 0 to 2·5pF), conductance from 0 to 100mmhos (lowest f.s.d. 0·01mmhos), inductance from 0·4μH to 100mH and reactance from 4Ω to 1MΩ. All measurements may be made over the frequency range of 15kc/s to 15Mc/s, the resistance, capacitance and conductance ranges being independent of frequency and the inductance and reactance calibrations applying where $\omega = 10^7$.

The sloping front panel carries two calibrated dials, the two range switches, the zero set controls and two coaxial sockets for the source and detector. The top platform carries six sockets comprising, earth, neutral, common and three ranges, for connexion of the apparatus or component under test.



Plug-in spring loaded terminals are provided which permit easy connexion. The plug-in feature is very useful in that it permits rapid interchange of jigs for production testing. Furthermore, the transistor adaptors available as an accessory can also be directly plugged in.

The bridge may be used with any suitable source or detector. However, where it is convenient to work at a fixed frequency, the source and detector assembly type LE.302, having $\omega = 10^7$, has been specially developed for use with the bridge, thus enabling it to be directly calibrated in inductance. Where this unit is fitted the detector takes the form of a milliammeter which is mounted between the two dials as illustrated.

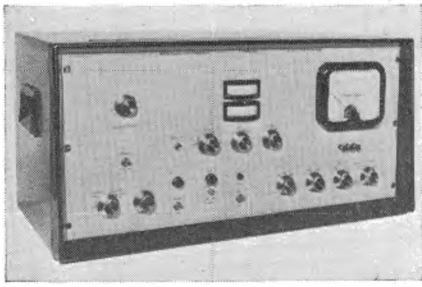
EE 24 762 *for further details*

TRANSISTOR TEST SET

Microcell Ltd, Blackwater Industrial Estate, Camberley, Surrey

(Illustrated above right)

This test set is designed to measure to a high degree of accuracy the parameters of pnp and npn transistors. The instrument incorporates several features which will commend it to the circuit designer who requires more accurate information about individual transistors than the manufacturers normally supply.



It is particularly suitable for the selection of matched sets of transistors.

Parameters which can be measured are amplification factor, cut-off frequency, collector leakage, and collector turnover voltage of transistors (grounded emitter configuration) and turnover voltage and leakage current of diodes.

Features of the instrument are the use of a differential input, wide-band valve-voltmeter as an amplification factor indicator and a wide-band oscillator as a signal source. The valve-voltmeter form of indicator effectively nullifies errors due to spurious voltages developed across the ratio networks. The response of this valve-voltmeter is within about 1dB from 1kc/s to 10Mc/s. The oscillator covers the range 1kc/s to 10Mc/s; it produces a stable output within +0 −1dB over this range. A bridge type control circuit ensures that output variation, with changes in ambient temperature is minimized.

Collector supplies—2 to 10V at 3A and 0 to 10V and 0 to 100V at 30mA—are fully stabilized as is the 220V, 200mA supply for the oscillator and valve-voltmeter.

Collector voltage and collector current are monitored on edge-type meters stacked one above the other so that both may be watched simultaneously. All meters are fully protected against overload. Provision is made for external monitoring of the base current.

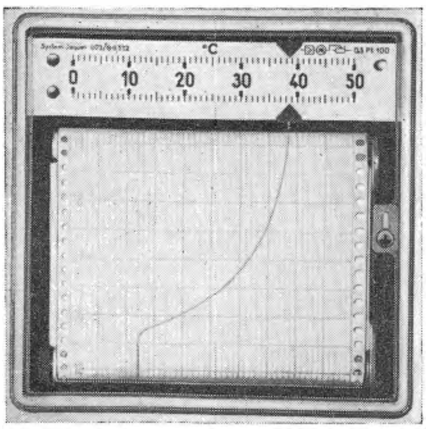
EE 24 763 *for further details*

POTENTIOMETER RECORDER

W. H. Joens & Co. GmbH, Martinstrasse 55, Düsseldorf

(Illustrated below)

This continuous-line recorder is housed in a case measuring 5⅜in × 5⅜in and has a recording width of 4in. The instrument is fully transistorized and in the interests of reliability and ease of maintenance electrolytic capacitors are not used and a fully temperature-compensated semiconductor constant voltage source is used instead of a standard cell.



The lowest measuring range is 5mV or 10μA d.c., and the speed of response 1sec f.s.d. The standard design includes changeover gears for eight chart speeds between 10 and 1 200mm/h and a voltage selector for six different voltages between 24 and 250V. A second slide wire and an adjustable set point potentiometer are incorporated to form an output bridge circuit for the representation of the set point deviation.

Various encapsulated plug-in units for the input circuit and the amplifier are available, according to requirements.

EE 24 764 *for further details*

MACHINE TOOL CONTROL EQUIPMENT

The General Electric Co. Ltd, Magnet House, Kingsway, London W.C.2

A new range of machine tool position and sequence control equipments is now available from the Electronics Division of The General Electric Company Ltd.

Both systems make use of a series of 'logic boards' and use transistorized printed circuits and cold-cathode valves. They are considerably smaller than conventional equipments, and the virtual elimination of electromechanical control gear reduces maintenance to a nominal routine electrical check.

The position sensing is by rotary digitizers which provide digital signal outputs representing the position of the work-table in two directions, and control the lead screw driving motors. Two-speed control is provided; a high rate of 90in/min for quick traverse and a low speed up to 12in/min for homing on to the required position with a final accuracy of ±0·0005in. This equipment is already successfully in use on the Newall 2436 Jig Borer.

Sequence control equipment provides both timing and triggering facilities for a multiple number of simultaneous and successive operations.

Installations to date include a vertical broaching machine and a medium sized injection moulding machine.

EE 24 765 *for further details*

FIELD PLOTTER

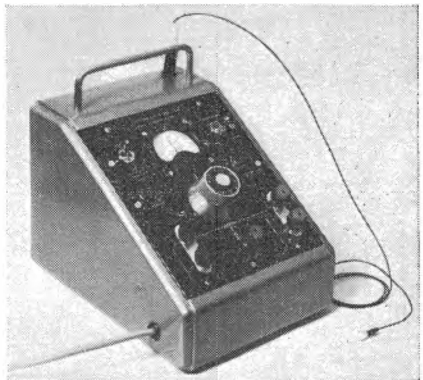
Servomex Controls Ltd, Crowborough, Sussex

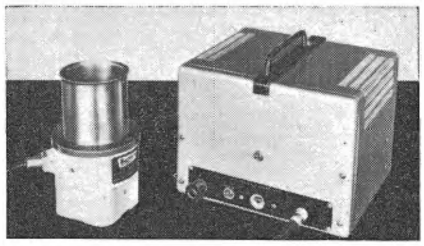
(Illustrated above right)

This new resistance analogue is really a two-dimensional version of the electrolytic tank and by means of the plotter the operator can draw a map of potential and flow lines conveniently and with no particular skill.

In use, the operator considers his problem and converts the quantities to electric field terms. Conducting electrodes are painted in silver paint in any desired shape on a plane of Teledeltos paper. The electrodes are connected to the FP.92 and ratio readings set on an accurately calibrated potentiometer. The ratio potentiometer tap connects with the probe through a galvanometer which indicates balance. The operator searches the field for balance which is marked with the pencil point on the probe. In a very few minutes sufficient equipotential lines can be plotted to give a graphic and accurate solution to the problem. Results are generally accurate to 5 per cent and can be closer with special effort.

EE 24 766 *for further details*





ULTRASONIC CLEANING UNITS

Kerry's (Ultrasonics) Ltd, Warton Road, Stratford, London E.15

(Illustrated above)

This 'Junior' range of ultrasonic cleaning units is suitable for cleaning small components such as ball bearing and ball race parts, instrument movements and parts, jewels, watch, clock and chronometer parts, thermionic and hydraulic valve components, germanium and silicon transistor and rectifier components, jewellery in manufacture, optical assemblies including lens frames, lenses, etc.

In addition to these components and many others the 'Junior' range of ultrasonic cleaners is suitable for use as a laboratory tool for evaluation work into ultrasonic techniques.

The units operate at a frequency of 40kc/s with the transducers mounted into the base of a round stainless steel container. The treatment container is mounted on a skirt or plinth designed to enable the unit to either stand on the bench or be positioned into a conventional type watch cleaning machine to form part of a normal cleaning sequence or cycle including rough wash, rinse and dry.

EE 24 767 *for further details*

BOOK REVIEWS

Modern Network Analysis

By F. M. Reza & S. Seely. 373 pp. 300 Figs. Medium 8vo. McGraw-Hill Publishing Co. Ltd. New York. London. 1959. Price 77s. 6d.

UNTIL the recent publication of Seshu and Balabanian's *Linear Network Analysis,* there had been an urgent need for a comprehensive exposition of modern network theory. Professor Reza of Syracuse University and Professor Seely of the Case Institute of Technology have made a further contribution to this need in their new book *Modern Network Analysis.*

The book is presented in three parts bound in one volume. Part 1 deals with the fundamentals of network analysis, Part 2 with two-port theory, and Part 3 is devoted to transient response and system functions. Part 1 contains five chapters, the first of which introduces the postulates of electric network theory. Because a large class of forcing functions encountered in network studies are *e*-type functions for which the methods of the Laplace transform are superfluous, the authors develop the s-plane representation of network behaviour with the aid of Campbell's cisoidal function. This is the topic of Chapter 2. The Laplace integral is not considered until later in the book. The remaining three chapters of Part 1 are concerned respectively with matrix algebra, network topology, and the general matrix theory of linear networks. The treatment of these subjects is particularly good.

Sufficient linear graph topology is covered in the text to enable the reader to set up the equilibrium equation of a given network in the most efficient way and an appendix is added at the end of the book in which the more sophisticated aspects of algebraic topology are presented. The terms employed in network topology, such as node, branch, linear graph, tree, tree-complement, cut-set, and tie-set are defined and the incidence matrix representation of linear graphs is discussed in detail. Topological duality is also considered. The essence of duality in network topology consists in the duality between cut-set and tie-set but these terms are defined in such a way that this duality is not evident. If a cut-set is defined as any set of branches in a linear graph which forms a single open parallel system when all branches not in the set are contracted, then the definition of a tie-set follows at once as any set of branches which forms a single closed series system when all branches not in the set are removed.

In the discussion of duality, the authors give a set of topological rules for finding the dual of an oriented planar linear graph. If these rules are applied to the dual thus obtained, the result is the original linear graph with all branch orientations reversed. This is surely incompatible with the concept of duality. As this reviewer has suggested elsewhere, since a planar linear graph is mappable on a simply-connected surface without the crossing of branches and since such a surface is two-sided, it seems reasonable to regard these two sides as duals. If the rules are applied to the graph as seen from the other side of the surface, a subsequent application of the rules yields the original graph properly oriented. It is significant that the two Kuratowski graphs, for which no duals exist, have to be mapped on a doubly-connected surface to avoid the crossing of branches and such a surface is one-sided!

Part 2 contains three chapters in which the six two-port matrices are derived and their use in the study of interconnected two-ports explained. Also considered are the two-port image parameters, conventional filter theory, and distributed-parameter two-ports. Part 3 contains five chapters and is devoted to a study of initially relaxed networks in the transient state. A form of Heaviside's second expansion theorem is derived and network stability is discussed. The basis of the treatment up to this point in the book is the cisoidal function but in the eleventh chapter the singularity functions are introduced and their use in the specification of discontinuous forcing functions is explained. In the following chapter the Laplace transform method of solving network integro-differential equations is briefly reviewed and related to the earlier parts of the book. The penultimate chapter deals with the natural modes of a linear network and the book concludes with a chapter on the analytical properties of physically realizable network functions.

Associated with each chapter are many examples showing how the various methods of analysis are applied and a wide variety of problems to be solved by the reader. The book does not take the subject quite so far as that of Seshu and Balabanian but it can be thoroughly recommended to any student wishing to acquire the necessary background to keep abreast of current developments in network theory.

S. R. DEARDS

The Cathode Ray Tube and its Applications

By G. Parr & O. H. Davie. 433 pp. 357 Figs. Demy 8vo. 3rd Edition. Chapman & Hall Ltd. 1959. Price 50s.

THE authors intended this revised edition to be a guide to the operation and uses of the cathode-ray tube in oscillography, and in this they have succeeded. The book has been split into approximately two halves one discussing the circuits associated with the c.r.t., which is well described, the other half discussing the applications of the c.r.t. However, so many applications have been mentioned that a number of the individual descriptions are of necessity very sketchy. The general style and presentation of the subject matter makes this a very readable book, the whole volume is well indexed, the paper of good quality and the circuit diagrams are clear and well defined. Each chapter is complete in itself and discusses first the general principles involved and then some practical solutions, followed by a very comprehensive bibliography.

The book appears to be primarily intended for the worker in an industrial applications laboratory. It is also of interest to the oscillograph designer as a guide to the variety of uses the instrument is asked to meet. The vast number of references also makes it useful for the serious student.

The first two chapters deal with the evolution of the cathode-ray tube to the present day, and describe the construction and performance of current, including p.d.a. types, of c.r.t. They also discuss in a general way the distortions commonly encountered. Then follow four chapters on circuits associated with the c.r.t., power supplies, amplifiers, linear time bases and other time bases. In each chapter the basic principles are encountered and individual circuits described. These circuit descriptions are very comprehensive but each one has tended to be treated in isolation. It would have been helpful to describe a complete amplifier and time base at the end of each relevant chapter. A useful chapter on auxiliary circuits describes some methods of calibrating the oscilloscope to measure time and voltage correctly. Also the basic design features of some fast rise time pulse generators are well described. Then follows a chapter on frequency boxes, which appears to cover all the current methods of swept frequency response curve plotting. The rest of the book, approximately half the total, is devoted to applications, and these cover a very wide range. A detailed chapter on photographic recording gives some very good practical advice and includes a short description of processing techniques. A description of a variety of transducers follows, which is too brief and warrants rather more detailed information on how they are used in practice. Three chapters on mechanical, electrical and radio engineering applications follow, which are very well described. The applications vary from the measurements of torque to a schematic diagram of an automatic valve tester. A chapter on television measurements briefly mentions some current testing techniques but is too compressed to do any justice to such a vast subject, although the 35 references are very useful. Again a chapter on general and nuclear physics does no more than scratch the surface, although the applications chosen are of great interest to the non-specialist. The last chapter on miscellaneous, including electro-medical, applications, describes the part played by the cathode-ray tube in some large scale systems.

L. D. KREPS

Electron Physics

By O. Klemperer. 220 pp. 80 Figs. Demy 8vo. Academic Press Inc., New York. Butterworths Scientific Publications, London. 1959. Price 32s. 6d.

IN this treatment of the basic physics of the free electron, the following topics are discussed: electron motion in vacuum, including space charge effects, and the production and detection of cathode rays: the charge and mass of the electron, and the relativistic change of the latter: the wave nature of the electron, and electron spin and magnetic moment. M.K.S. units are used, and the standard of presentation is high.

The book is intended primarily for undergraduates, and is to be welcomed because it cuts across the outdated academic boundaries which divide physics as it is usually taught. Because the treatment is so successful, the book will appeal to a much wider readership, for few will not discover something new in it, or fail to be stimulated by the juxtaposition of even the most familiar ideas. For the electronic engineer here is a valuable refresher course in fundamentals, at a time when few electronic processes are not being developed into devices of potential practical usefulness.

Particularly refreshing is the brevity of the book, stemming as it does from concise argument and economy of words. Unfortunately the rigorous compression necessary has led to some distortion on emphasis. Valuable though the discussion of the measurement of e and e/m may be, it inevitably occupies a disproportionate amount of space in a book of this length, compared with the more important matters of relativistic effect or the wave properties of the electron. In particular one wishes that the latter topic could have been expanded to set the reader on the threshold of gaseous and solid state electronics, yet this remains a minor blemish on a valuable book.

G. F. ALFREY

Analytical Transients

By T. C. G. Wagner. 202 pp. 105 Figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd. London. 1959. Price 70s.

THIS slim volume is addressed to the reader who has previously studied the elements of network analysis, the differential and integral calculus, and linear algebra, but who may not have encountered Fourier series and the Laplace transformation. The purpose of the book is to provide the more advanced mathematical knowledge necessary for a fuller understanding of linear network behaviour.

The book contains seven chapters and begins with an introduction to the Laplace integral and Heaviside's second expansion theorem. This is followed by a description of the various methods of setting up the equilibrium equations of a given network of linear electrical or mechanical elements. The topics discussed include linear graph theory, Kirchhoff's laws, d'Alembert's principle, and Lagrange's equations. The influence of discontinuities in the forcing function upon the response of a system is considered next. Several pages are devoted to a method of determining the response and its derivatives immediately after a discontinuity in the forcing function; the method is illustrated with several examples. Fourier analysis is then introduced. Among the subjects discussed in this connexion are Parseval's theorem, convergence, the Fourier integral and its relation to the Laplace integral, periodic and non-periodic functions, and Gibbs' phenomenon. The stability theorems of Sturm, Cauchy, Nyquist, and Routh are the topics of the fifth chapter and the last two chapters are concerned with the Laplace transformation and complex variable theory.

As stated in the preface, incomplete logical arguments and holes in mathematical proofs have been avoided. The chapter dealing with the effect upon the response function of discontinuous forcing functions contains some new material. The author holds that the definitions of the network elements and Kirchhoff's laws are of themselves sufficient to determine the response of any network to a transient and that the concepts of conservation of flux-linkage and charge are dependent upon the time invariance of the network equations.

This is a book for the serious student.

S. R. DEARDS

Electronic Computers—Principles and Applications

By T. E. Ivall. 263 pp. 56 Figs. Demy 8vo. 2nd Edition. Iliffe & Sons Ltd. 1960. Price 25s.

DEVELOPMENTS since the first edition was published in 1956 have been so rapid that the present volume has had to be almost entirely rewritten, most of the existing diagrams replaced by new ones, and pictures of more modern electronic computer equipment introduced. Three new chapters have also been added, dealing respectively with analogue computer circuits, the programming of digital computers and the evolution of the more 'intelligent' machines of the future.

The fundamental approach to the subject has, however, remained unchanged and as before the book is intended as a non-mathematical introduction to the principles and applications of computers employing valves, transistors, and other electronic devices; it has been designed to appeal to technicians, engineers and students who have some knowledge of electronic or electrical engineering, but some chapters are also suitable for the interested layman.

The Technical Writer

By J. W. Godfrey and G. Parr. 333 pp. 106 Figs. Demy 8vo. 2nd Edition. Chapman & Hall Ltd. London. 1960. Price 45s.

Few editors would dispute the general statement that the standard of composition and basic preparation of the majority of manuscripts they receive is appallingly low. This book is splendidly written, it is most comprehensive and no author could fail to benefit from its teachings. It is to be feared, however, that its length and price will tend to keep it from the eyes of the casual author and one would need to be an optimist indeed to hope that it will help directly to raise the present standards.

This is a comprehensive text-book designed to give the technical writer a broad insight not only into the basic principles of good writing but also into the whole technique of presentation and production. Let it be hoped that means will be found to distil at least some of its teachings into the minds of all potential authors. The standard of their manuscripts would be greatly raised and the task of editors proportionately eased.

G. R. M. GARRATT

Elements of Radio Engineering

By H. L. F. Peel. 254 pp. 158 Figs. Crown 8vo. 2nd Edition. Cleaver-Hume Press Ltd. 1960. Price 13s. 6d.

THE revisions for this new edition are improvements of detail rather than form and the general scope and arrangement have been maintained. The most notable amendments and additions are those relating to static and dynamic characteristics, transformer cooling, lay-out of receivers and amplifiers, power factor, and transistors. The selection of examination questions given on each chapter has been revised and enlarged in accordance with recent trends.

The book assumes only an elementary knowledge of electricity and magnetism, and develops simple a.c. theory from the radio engineer's angle.

V.H.F. Line Techniques

By C. S. Gledhill. 57 pp. 50 Figs. Demy 8vo. Edward Arnold (Publishers) Ltd. 1960. 12s. 6d.

THIS book has been written with the undergraduate student primarily in mind, for it deals in some detail with the various methods of using transmission lines as elements to perform various functions of communication circuits. Wherever applicable, the rigorous analysis of the method is given, but the aim in every section has been to show that the circle-diagram technique yields the same results as the analysis and also does it more rapidly. Throughout the book are worked examples of a practical nature, which will be helpful to the practising engineer as well as the student.

La Science et la Théorie de l'Information

By L. Brillouin. 302 pp. 74 Figs. Medium 8vo. Masson et Cie, Paris. 1960. Price 4.800fr.

PROFESSOR LEON BRILLOUIN is well known and a book from his pen should be received with both enthusiasm and respect. There is no doubt that this book is worthy of its author.

The first quarter of the book is devoted to the definition of information and its application to particular problems such as redundancy, coding and classification. This is followed by the definitions and basic notions necessary for the comprehension of the following chapters. Here the author has carefully avoided overloading by restricting himself to what is absolutely necessary and where what may appear to be categorical statements are made, they are backed up by an extensive bibliography.

In the main body of the work appears the author's remarkable correlation between information and entropy which leads him to the double conclusion that information is in fact entropy with a simple change of sign, for which he has coined the word "neguentropie", and that any experiment which supplies information, inevitably results in an increase of entropy. The application of this conclusion to a number of interesting special cases clarifies several problems of which the real significance has often remained obscure for engineers and even for some physicists.

In this book there is the rare combination of a text which is easy to read and one which stimulates reflection. It ought to be in the library not only of those concerned with telecommunications and computers but of physicists and engineers whose work is connected in any way whatsoever with measurement.

G. Ashdown

Principles of Servomechanisms

By A. Tyers and R. B. Miles. 173 pp. 182 Figs. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1960. Price 25s.

IN as far as this book is an 'appetizer' and does not really deal with the 'main course' of the subject one feels that the title is misleading. The more so for the old hands of the subject of servomechanisms who will recall perhaps with nostalgia that over 12 years ago the first really comprehensive pedagogic text on the subject bore the same title. The present volume is not (and neither did its authors intend it to be), in the same class. It is intended for junior development and maintenance engineers with no mathematical knowledge beyond elementary trigonometry but a familiarity with general electronic circuits. As such it would have been more happily entitled an 'Introduction'.

To deal with servo theory without using differential equations means that the subject must of necessity be dealt with descriptively and can only be qualitative rather than quantitative. Given these restrictions the authors are ingenious in surmounting their chosen limitations and do provide in an entertaining way an introduction to the jargon of the subject. Since a.c. vectors are used in describing the action of a.c. servos and synchros one wonders why the authors saw fit to omit all reference to frequency response, the major hurdle having been taken.

The major part of the book is concerned with the functioning of the peripheral hardware; error detectors, modulators and demodulator amplifiers (d.c., a.c. magnetic, thermionic, transistor and thyratron), servo motors and tachometers; topped-off for good measure by a chapter on analogue computers (again without resort to differential equations).

The authors have set out to help junior engineers and technicians who find themselves involved with servomechanisms to understand in general terms what goes on in them. Most electrical components and circuits peculiar to servos are described and practical details of their construction given. The explanations are well matched to the readers they are intended for and will be readily understood by them. This is an important point in a book of this sort and the authors have shown real understanding of the need for absolute clarity. As an 'appetizer' the book is well-conceived and recommendations are given at the end for wider reading for those who stay the course with an appetite for a rather heavier 'course'.

J. H. Westcott

The Instrument Manual

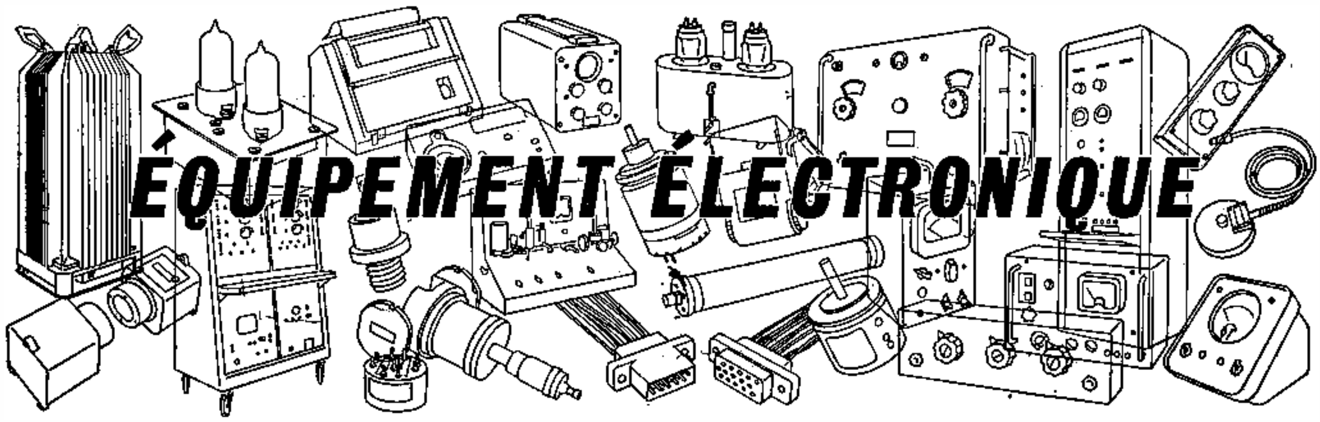
742 pp. 998 Figs. Demy 4to. 3rd Edition. United Trade Press Ltd. 1960. Price 105s.

THE general scheme of this edition follows that of the second edition. One section has been omitted, namely Navigational Instruments, and this has been replaced by one on Analogue Computers. The original optic section has been dispersed. Microscopy has now been given a section to itself, and the analytical pattern of optical instruments has been included in the section on Analytical Instruments, which also includes all relevant material from the section in the second edition called Instruments for the Determination of Compositional Quality. The titles of some of the sections have been modified to indicate their contents more accurately.

Theory and Design of Inductance Coils

By V. G. Welsby. 226 pp. 57 Figs. Demy 8vo. 2nd Edition. MacDonald & Co. (Publishers) Ltd. 1960. Price 30s.

THE material of the first edition has been completely revised and new matter added. Resulting from the development of microwave radio and of h.f. techniques during recent years, much interest has been created in the principles of electromagnetic wave propagation. In this second edition this aspect is referred to and it is shown how a picture derived from electromagnetic field theory simplifies the understanding of inductors. In this new edition the units are in the modernized m.k.s. system.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 516 à 519

CALCULATEUR ANALOGIQUE "SUR CHAÎNE"

(Illustration à la page 516)

De Havilland Propellers Ltd, Hatfield Aerodrome, Hatfield, Hertfordshire

Le calculateur analogique "ANATROL" constitue une conception nouvelle dans l'étude des systèmes de calcul analogique "sur chaîne" et fournit, pour la première fois, un calculateur bon marché, complet et adaptable, ayant la robustesse et la stabilité indispensables pour le contrôle moderne des procédés de fabrication à la chaîne.

La précision de calcul d'ANATROL est compatible avec le degré de précision qu'on exige actuellement des mesures, ce qui lui permet de s'intégrer facilement et efficacement à des dispositifs de contrôle existants.

Les calculateurs analogiques classiques fournissent les solutions d'équations de procédés à des vitesses extrêmement rapides, comparées à la dynamiques des procédés. Ce fait, allié à la nécessité d'augmenter la sûreté de fonctionnement et de réduire considérablement le prix, a conduit au choix des principes utilisés.

Les minima des amplificateurs et multiplicateurs de calcul sont formés en unités arithmétiques. Ces minima sont ensuite employés pour apporter une solution point par point des équations, selon un programme pré-établi dans lequel chaque point est choisi en séquence par un système de commutation. Des petites matrices à mémoire sont employées pour retenir les informations sur le résultat d'un point dans la solution, ces données étant utilisées à un stade ultérieur du calcul. De longues matrices sont employées pour fixer les sorties sur le dernier chiffre calculé, cependant qu'on détermine de nouvelles valeurs.

Un maximum de vingt-cinq points est prévu, ce qui donne à la machine le potentiel d'un calculateur employant au moins soixante quinze amplificateurs de calcul et vingt-cinq multiplicateurs. Des méthodes à base de semi-conducteurs sont appliquées exclusivement, à l'exception de quelques éléments de commutation électro-mécaniques. La machine est de construction modulaire, et différentes combinaisons composante-unité peuvent être employées selon le problème traité. Au besoin, un élément supplémentaire d'entrée/sortie peut être fourni. Ce dernier contient des transducteurs ou amplificateurs, pour adapter le calculateur exactement à l'équipement de mesure et de contrôle de l'usine.

EE 24 751 *pour plus amples renseignements*

CONTRÔLEUR DE LIMITES DE VIBRATIONS

(Illustration à la page 516)

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

Le contrôleur de limites de vibrations type 1434 a été créé pour mesurer, contrôler et enregistrer les vibrations dans les machines et les structures. Le contrôle de vibrations excessives des grands ensembles de machines rotatives, comprenant du matériel hydroélectrique et turboélectrique, ainsi que les ponts et barrages lourdement chargés, peut se faire automatiquement en comparant continûment le signal provenant d'un capteur à une tension de polarisation préréglée. Un tube de déclenchement avec un relais de grande puissance dans le circuit anodique est déclenché si le signal est supérieur au niveau prédeterminé. Le relais arrête alors la machine ou actionne un dispositif d'alarme.

Le capteur de vibrations est de construction robuste, à cadre mobile. Il est hermétiquement scellé pour maintenir la précision dans les conditions de température et d'humidité élevées. Des conducteurs jumelés à gaine, ayant jusqu'à 200 pieds de long, permettent de placer le contrôleur à distance sûre des capteurs montés sur des machines en des endroits dangereux ou inaccessibles. Dans la gamme de fréquences allant de 20 à 400 Hz (1 200 à 24 000 tours/minute), le capteur fournit une tension de sortie proportionnelle à la vitesse de vibration de la surface avec laquelle il est en contact. Un signal représentant le déplacement s'obtient en appliquant la tension de sortie à un montage intégrateur, suivi par l'amplification et le redressement. Le signal c.c. qui en résulte est utilisé pour actionner un appareil de mesure étalonné en valeurs de pointe de déplacement pour un simple mouvement harmonique appliqué au capteur. Cette indication du niveau moyen de vibration, donnée par l'appareil de lecture, peut être enregistrée en enfichant un milliampèremètre enregistreur c.c. ordinaire.

Une commande de sensibilité offre quatre gammes, soit: 0 à 0,001 pouces, 0 à 0,003 pouces, 0 à 0,010 pouces et 0 à 0,030 pouces (unités anglaises) ou 0 à 0,03 mm, 0 à 0,1 mm, 0 à 0,3 mm et 0 à 1 mm (unités métriques). Le contrôleur peut être réglé pour un declenchement à n'importe quel niveau de vibration entre 20 et 100% de déviation totale de l'appareil de mesure, dans n'importe quelle gamme.

EE 24 752 *pour plus amples renseignements*

ANALYSEUR DE FRÉQUENCES

(Illustration à la page 516)

Laboratoire Electro-Acoustique, 5, rue Jules-Parent, Rueil (Seine-et-Oise), France

Cet instrument a été réalisé pour l'analyse des composantes d'un spectre de fréquences ayant une fréquence fondamentale de 50 Hz.

L'analyse s'effectue de la façon suivante:

Une fréquence variable, produite par un générateur de précision incorporé à l'analyseur, est superposée au spectre devant être analysé. Après détection, on attaque un amplificateur sélectif, avec appareil de mesure, qui ne permet que le passage d'une bande étroite. Chaque fois que la différence entre la fréquence variable et la composante d'un spectre donne une fréquence dans la bande passante, on obtiendra une déviation de l'instrument, proportionelle à l'amplitude de la composante.

L'analyseur F.A.H. 17 permet aussi bien l'examen des composantes, l'une après l'autre, que celui d'une composante individuelle, en fonction du temps.

Etant donné que la fréquence fondamentale de 50 Hz peut ne pas être absolument stable, il est nécessaire d'avoir une bande passante d'une certaine largeur, afin d'assurer l'accord des différentes composantes. En effet, si l'accord est trop aigu, des variations de

(Pages 523 to 532 French and German Sections)

fréquence, même fort limitées, causeraient des variations dans l'indication de l'instrument, sans qu'il n'y ait, de fait, une variation dans la tension de la composante.

D'autre part, il est indispensable que la sélectivité de l'instrument, en dehors de la susdite bande passante, soit suffisamment grande pour empêcher que la mesure d'une composante en cours d'examen, ne soit influencée par des composantes voisines, de chaque côté de celle-là.

La gamme de fréquences de l'analyseur est de 100 Hz à 2 kHz, et l'affaiblissement est de 25 dB à ∓50 Hz, à partir du milieu de la bande passante. Une sortie est prévue pour l'utilisation d'un enregistreur à plume.

EE 24 753 *pour plus amples renseignements*

POTENTIOMÈTRES MINIATURE

(Illustration à la page 517)

Miniature Electronic Components Ltd. The Lye, St. Johns, Woking, Surrey

La susdite société fabrique sous licence Con-Elco de Californie, une série de potentiomètres d'équilibrage miniature, livrables dans les trois modèles courants qu'on voit sur notre gravure.

La gamme de températures va de − 55° C à + 180° C, la puissance nominale est de $2\frac{1}{4}$ W et la valeur de décroissement nominal va jusqu'à $\frac{1}{2}$ W à 150° C. La gamme de résistance s'étend de 10 Ω à 50 kΩ. Ces potentiomètres ont été spécialement conçus et réalisés pour l'industrie des engins téléguidés, où prévalent des conditions d'environnement fort rigoureuses et où la sûreté de fonctionnement est des plus importantes.

EE 24 754 *pour plus amples renseignements*

COMPTEUR DE PUISSANCE B.F.

(Illustration à la page 517)

Amos of Exeter Ltd, Weircliffe Court, Exwick, Exeter, Devonshire

La conception du modèle 157 se rapproche sensiblement de celle des autres compteurs de puissance BF Weircliffe utilisant un mouvement à cadre mobile de 114 mm, à échelle de 0 à 10 mW et − 20 dB à + 10 dB. Toutes les résistances de charge, bobinées inductivement et d'une puissance nominale à l'épreuve des conditions continues de pleine charge, sont réglées pour une précision de 0,1%. Un réglage variable indépendant peut être effectué pour compenser les variations individuelles dans les mouvements de compteur.

L'accord des impédances d'entrée se fait par un transformateur à noyau de type "C" de dimensions convenables ayant un noyau en bande de 0,1 mm afin d'assurer une réponse de fréquence linéaire. Une prise médiane autorise une entrée pleinement équilibrée à toutes les impédances et l'isolement est tel que l'instrument peut être relié directement à un circuit à lampes pour déterminer les conditions de charge optimum. Les deux commandes d'impédances fondamentales choisissent des valeurs de 1 à 10 ohms par paliers de 0,5 ohm, tandis qu'un multiplicateur donne des facteurs de X1, X2, X5, X10, X20, X50, X100, X200, X500, et X1 000, permettant un choix rapide de n'importe quelle impédance de 1Ω à 10kΩ.

Un commutateur à cinq directions, directement étalonné, choisit des gammes de 10 mW, 100 mW, 1 W, 10 W et 50 W de déviation totale, la lecture utile la plus basse dans chaque gamme étant un dixième de l'échelle totale.

EE 24 755 *pour plus amples renseignements*

ENREGISTREUR À DOUZE CANAUX

(Illustration à la page 517)

AB Elektrisk Malmletning, Danderydsgatan 11, Stockholm

Les enregistreurs à plume à écriture directe de la série ABEM constituent des instruments de précision pour l'enregistrement d'un à douze phénomènes transitoires à vitesse réduite. Ces derniers peuvent être injectés aux galvanomètres à cadre mobile à prise médiane, soit comme signaux c.a. soit comme signaux c.c. L'enregistrement direct s'obtient par l'emploi de plumes chauffées et de papier ciré. Vu que le papier passe sous les plumes, la surface cirée en contact avec celles-ci fond, faisant ainsi apparaître le fond de couleur rouge vif du papier, qui produit une trace fortement contrastée. Deux marqueurs de phénomènes transitoires sont fournis, permettant d'enregistrer le temps ou des repères d'identification.

En échangeant les roues d'engrenage, on peut obtenir trois vitesses complémentaires dans la gamme de 1 mm/min. à 10 cm/sec. Il y a, en outre, un choix de galvanomètres avec résistances intérieures de 1750 Ω, 430 Ω, 100 Ω, 28 Ω, avec sensibilités de 25 mm/1 mA, 25 mm/2 mA, 25 mm/4 mA et 25 mm/8 mA respectivement.

EE 24 756 *pour plus amples renseignements*

OSCILLATEUR KLYSTRON RÉFLEX

(Illustration à la page 517)

English Electric Valve Co. Ltd, Chelmsford, Essex

Le nouveau klystron réflex basse tension K.361 a été spécifiquement étudié à l'intention des systèmes Doppler de mesure de vitesse, pour lesquels un fonctionnement sûr avec un minimum de soins routiniers sont des conditions essentielles.

Des applications typiques comprennent un équipement produit par la Westinghouse Brake & Signal Co. pour la commande automatique de vitesse de matériel roulant garé dans des voies de classement et de systèmes Doppler de "radars" de police pour les mesures de vitesses. L'accord mécanique du klystron s'effectue, dans la gamme de fréquences de 10700 à 10725 MHz, par un bloc d'accord à vis unique pénétrant dans la cavité.

Le klystron réflex K.361 a une puissance de sortie minima de 20 mW et il est muni d'un coupleur type UG-40 A/U (Z830051) pour la connexion au guide d'onde No. 16.

EE 24 757 *pour plus amples renseignements*

BLOCS D'ALIMENTATION STABILISÉE

(Illustration à la page 517)

International Electronics Ltd, 132-135 Sloane Street, London S.W.1

Ces blocs entièrement transistorisés sont montés sur châssis d'acier zingué de 1,63 mm et peuvent être fixés sur panneau ou plaque de base. N'importe quels deux blocs peuvent être montés côte à côte sur panneau à support normal de 48,2 cm.

Dans leur assortiment actuel, ces blocs couvrent des courants de 100 mA, 200 mA, 350 mA et 500 mA. aux tensions normales de 200 V, 250 V et 300 V. En outre, tous les blocs sont munis de deux bobinages de chauffage, dont l'un est à prise médiane.

Les blocs sont pleinement protégés par des fusibles de secteur et de sortie.

L'impédance de sortie à 500 kHz est inférieure à 0,5 Ω, l'ondulation efficace est inférieure à 1 mV et le facteur de stabilité est de 400 : 1.

La première série de blocs transistorisés d'alimentation à basse tension vient également d'être lancée. Ces blocs fournissent un courant de 2 A à n'importe quelle tension entre 5 V et 50 V et ils utilisent une méthode de commutation qui évite l'emploi d'un grand nombre de transistors en parallèle. Des blocs fournissant des courants de 5 A et 10 A seront ajoutés prochainement à cette série.

EE 24 758 *pour plus amples renseignements*

POTENTIOMÈTRE RECTILINÉAIRE

(Illustration à la page 518)

General Controls Ltd, 13-15 Bowlers Croft, Honywood Road, Basildon, Essex

Ce potentiomètre rectilinéaire de précision, avec frotteurs à suspension équilibrée, a été realisé pour parer au problème de la séparation du frotteur de l'élément de résistance, dans des conditions environnantes extrêmes de chocs et de vibrations. Le potentiomètre a un frotteur en forme de "U", tenu en équilibre de pression entre la tige de commutation et l'élément résistif. Toute force extérieure tendant à modifier la pression de contact du frotteur contre l'élément résistif, provoquera une réaction opposée égale sur le côté de commutation du frotteur en "U", maintenant ainsi une pression constante du bras du frotteur. Ce système réduit au minimum le bruit de contact et la discontinuité du signal sélectif et empêche une pression nuisible du frotteur contre l'élément résistif.

Le nombre maximum de prises, y compris les connexions d'extrémité, est de cinq, et la gamme de résistances va de 100 Ω à 10 kΩ.

EE 24 759 *pour plus amples renseignements*

RADIOGONIOMÈTRE AUTOMATIQUE M.F.

(Illustration à la page 518)

Marconi's Wireless Telegraph Co. Ltd, Chelmsford, Essex

Le radiogoniomètre Type ND.103 comprend le système à cadres fixes Bellini-Tosi, et l'élément rotatif est constitué par une bobine de goniomètre automatiquement réglée, mise en action selon la méthode du signal nul.

Un récepteur sensible et un radiogoniomètre sont logés dans un seul bloc compact, propre au montage sur établi. Quoique l'appareil soit principalement prévu pour le fonctionnement automatique, il est aussi doté de commandes manuelles, sous une plaque à charnières, les commandes essentielles pour l'emploi automatique étant aisément accessibles et commodément placées pour l'utilisateur.

Le panneau avant est en pente, afin d'en faciliter la lecture, et il porte à droite l'échelle d'accord et à gauche l'échelle d'indication de relèvement, cependant que le compteur d'accord se trouve entre les échelles.

L'appareil peut être équipé de tous les principaux types de répéteurs gyroscopiques, évitant ainsi la nécessité de pointer le cap du navire, et il se conforme rigoureusement aux spécifications des services postaux britanniques, en ce qui concerne l'équipement radiogoniométrique pour navires.

La précision des relèvements sur "automatique" est supérieure à $\pm 1\%$ et un étalon intérieur à cristal assure la vérification d'échelle.

La commutation "d'automatique" à "manuel" peut s'effectuer immédiatement, ce qui permet de relever des signaux faibles en même temps que des signaux plus forts, cas qui peut se produire quand on répond à un signal de détresse.

EE 24 760 *pour plus amples renseignements*

BLINDAGE DE CHLORURE DE POLYVINYLE POUR TEMPÉRATURES ÉLEVÉES

Permanoid Ltd, New Islington, Manchester, 4

Pour les applications ou le traitement à températures élevées, la société Permanoid a créé un nouveau doublage de chlorure de polyvinyle avec renforcement intérieur de soie visqueuse. Ce matériel se conforme à la Norme Britannique B.S. 2848, 1957, Classe 105T, Type 1.

Le blindage à soie visqueuse Permanoid allie les avantages d'un noyau de textile intégral et d'une gaine en chlorure de polyvinyle sans aucun des désavantages. Le renforcement intérieur en soie visqueuse assure la flexibilité d'emploi et élimine le risque de rétrécissement durant la cuisson. Une couche extérieure en chlorure de polyvinyle plastifié augmente la densité et, par conséquent, la résistance aux huiles ou vernis d'imprégnation, réduisant ainsi efficacement la possibilité de panne mécanique durant le traitement. Il a de bonnes propriétés de vieillissement, il est imperméable à la moisissure tropicale et convient à des températures de fonctionnement continu allant jusqu'à 105°C.

Le blindage Permanoid est fourni en onze couleurs (noire et blanche y comprises) et dans des diamètres d'alésage de 0,5 mm à 5 mm.

EE 24 761 *pour plus amples renseignements*

PONT HF

(Illustration à la page 518)

Hatfield Instruments Ltd, Crawley Road, Horsham, Sussex

Le pont haute fréquence LE300 assure, en cinq gammes, l'étalonnage direct de résistances de 10 Ω à 10 M, de capacités de 0 à 0,025 µF (la gamme la plus réduite étant de 0 à 2.5 pF), de conductances de 0 à 100 mmhos (la déviation totale la plus réduite étant de 0,01 mmhos), d'inductances de 0,4 µH à 100 mH et de réactances de 4 Ω à 10 M. Toutes les mesures peuvent être effectueés dans la plage de fréquences de 15 kHz à 15 MHz, les gammes de résistances, de capacitances et de conductances étant indépendantes de la fréquence et les etalonnages d'inductances et de réactances appliquant $\omega = 10^7$.

Le panneau à avant en pente porte deux cadrans étalonnés, ainsi que les deux commutateurs de gammes, les commandes des réglage de zéro et deux prises coaxiales femelles, pour la source et le détecteur. Le plateau supérieur porte six prises, dont une de mise à la terre, une neutre, une commune et trois prises de gammes, pour relier l'appareil ou la pièce détachée soumis au contrôle.

Des bornes interchangeables à ressorts facilitent la connexion. Cette qualité d'interchangeabilité est extrêmement utile, car elle permet l'échange rapide des gabarits de contrôle de production. De plus, les adaptateurs de transistors, fournis comme accessoires, peuvent aussi être enfichés directement.

Le pont HF peut être utilisé avec n'importe quelle source ou n'importe quel détecteur appropriés. Lorsqu'il est plus commode, cependant, de travailler à une fréquence fixe, on utilisera l'ensemble de source et détecteur type LE.302, dont $\omega = 10^7$, et qui a été spécialement étudié pour le pont, permettant ainsi de l'étalonner directement en inductance.

EE 24 762 *pour plus amples renseignements*

APPAREIL DE CONTRÔLE À TRANSISTORS

(Illustration à la page 519)

Microcell Ltd, Blackwater Industrial Estate, Camberley, Surrey

Cet appareil de contrôle a été créé pour mesurer, à un degré élevé de précision, les paramètres de transistors pnp et npn. Il comporte plusieurs avantages qui le recommandent au réalisateur de circuits ayant besoin de plus de renseignements sur des transistors individuels que n'en fournissent habituellement les fabricants. Il est particulièrement indiqué pour le choix de jeux de transistors équilibrés.

Les paramètres pouvant être mesurés sont le facteur d'amplification, la fréquence de coupure, la puissance dissipée au collecteur, la tension de retournement au collecteur des transistors (montage émetteur à la masse), la tension de retournement et le courant de fuite des diodes.

L'appareil se caractérise par une entrée différentielle, un voltmètre à lampes à large bande servant d'indicateur de facteur d'amplification et un oscillateur à large bande faisant fonction de source de signaux. L'indicateur sous forme de voltmètre électronique élimine efficacement les erreurs dues aux tensions parasites se manifestant dans les réseaux de rapport. La réponse de ce voltmètre électronique est, à environ 1 dB près, de 1 kHz à 10 MHz. L'oscillateur couvre la gamme de 1 kHz à 10 MHz et produit une sortie stable, à + 0 − 1 dB près, dans cette gamme. Un circuit de commande, du type à pont, réduit au minimum les variations de sortie en fonction des changements de température ambiante.

Les alimentations au collecteur, soit 2 − 10 V à 3 A, 0 − 10 V et 0 − 100 V à 30 mA, sont pleinement stabilisées, de même que l'alimentation de 220 V, 200 mA. pour l'oscillateur et le voltmètre électronique.

La tension au collecteur et le courant au collecteur sont contrôlés par des indicateurs du type à bord, posés l'un sur l'autre, de manière à ce que les deux puissent être observés simultanément. Tous les indicateurs sont pleinement protégés par disjoncteurs à maxima. Le courant de base peut être contrôlé extrérieurement.

EE 24 763 *pour plus amples renseignements*

POTENTIOMÈTRE ENREGISTREUR

(Illustration à la page 519)

W. H. Joens & Co. GmbH, Martinstrasse 55, Düsseldorf

Ce potentiomètre enregistreur est logé dans un coffret de 144 mm × 144 mm et sa largeur d'enregistrement est de 100 mm. Il est entièrement transistorisé et ne fait pas usage de condensateurs électrolytiques, ce qui le rend d'un fonctionnement plus sûr et d'un entretien plus facile. De plus, une source de tension constante semi-conductrice, thermiquement réglée, est employée au lieu d'une pile étalon.

La gamme de mesure la plus basse est de 5 mV ou 10 µA c.c. et la vitesse de réponse est de 1 sec de déviation totale. Le modèle courant comprend des commutateurs pour huit vitesses de bande, de 10 à 1 200 mm/heure, et un sélecteur de tension pour six tensions différentes, de 24 à 250 V. Un second fil à contact glissant et un potentiomètre à point de repère réglable sont incorporés à l'enregistreur afin de former un montage en pont de sortie pour représenter la déviation du point de repère.

EE 24 764 *pour plus amples renseignements*

ÉQUIPEMENT DE CONTRÔLE DE MACHINES-OUTILS

The General Electric Co. Ltd, Magnet House, Kingsway, London W.C.2

Une nouvelle gamme d'appareils de contrôle de séquences et de la position de machine-outils est maintenant offerte par la Division d'Électronique de la General Electric Co. Ltd.

Les deux systèmes comportent une série de "tableaux logiques" et ils utilisent des circuits imprimés transistorisés et des lampes à cathodes froides. Ils sont considérablement plus petits que les appareils ordinaires. D'autre part, l'élimination virtuelle du matériel de commande électromécanique réduit l'entretien à une vérification électrique routinière.

La détection de position se fait par palpeurs à conversion numérique rotatifs, qui fournissent les sorties de signaux numériques représentant la position de la table porte-pièce dans deux directions et commandent les moteurs d'entraînement à vis mère. Une commande à deux vitesses est prévue, soit une grande vitesse de 2,3 m/min. pour un balayage rapide, et une vitesse réduisible jusqu'à 305 mm/ min. pour le retour à la position requise, avec une précision finale de ± 0,0127 mm. Cet équipement est déjà en service avec plein succès sur l'aléseuse Newall 2436.

L'équipement de contrôle de séquences permet le minutage et le déclenchement d'un nombre multiple d'opérations simultanées ou successives.

EE 24 765 *pour plus amples renseignements*

ABAQUE DE CHAMP

(*Illustration à la page 519*)

Servomex Controls Ltd, Crowborough, Sussex

Ce nouvel instrument analogique à résistance, est en réalité une version bidimensionnelle de la cuve rhéographique et, grâce à l'abaque, l'opérateur peut tracer une carte des lignes de potentiel et de débit, commodément et sans habileté particulière.

L'opérateur étudie le problème à résoudre et convertit les quantités en termes de champ électrique. Des électrodes conductrices sont peintes en couleur argent, en leur imprimant n'importe quelle forme voulue sur une surface de papier Télédeltos. Elles sont reliées au FP.92 et les lectures de rapport sont fixées sur un potentiomètre étalonné avec précision. La prise du potentiomètre de rapport est reliée au palpeur, à travers un galvanomètre indiquant l'équilibre. L'opérateur explore le champ pour trouver l'équilibre qui est marqué avec la pointe du crayon sur le palpeur. Au bout de quelques minutes, des lignes équipotentielles peuvent être tracées afin de donner une solution graphique et précise au problème.

EE 24 766 *pour plus amples renseignements*

APPAREILS DE NETTOYAGE À ULTRA-SONS

(*Illustration à la page 519*)

Kerry's (Ultrasonics) Ltd, Warton Road, Stratford, London E.15

Cette gamme d'appareils de nettoyage "Junior" à ultrasons est destinée au nettoyage de petites pièces telles que celles des roulements et cages à billes, de mouvements et pièces d'instruments, de rubis, de pièces de montres, horloges et chronomètres, de pièces de tubes thermo-électroniques et hydrauliques, de pièces de transistors et redresseurs au germanium et au silicium, de joaillerie, d'assemblages d'optique, tels que montures de lunettes, lentilles, etc.

En plus de son application à ces pieces et à une foule d'autres, la gamme des appareils à ultra-sons "Junior" se prête aussi à l'emploi comme outils de laboratoire, pour des travaux d'estimation de méthodes à ultra-sons.

Les appareils "Junior" fonctionnent à une fréquence de 40 kHz avec transducteurs montés sur la base d'un bac rond en acier inoxydable. Le bac de traitement est monté sur plinthe, pour que l'appareil puisse être érigé sur établi ou inséré dans une machine de nettoyage de montres de type ordinaire, afin qu'il fasse partie d'un cycle ou d'une séquence de nettoyage normaux comprenant le lavage, le rinçage et le séchage. Les pièces peuvent frequemment être nettoyées sans qu'il soit nécessaire de les demonter, ce qui évite les risques d'endommagement mécanique. Enfin, des films d'impuretés tenaces, tels que le rouge d'Angleterre, le tripoli ou terre pourrie, etc. peuvent être facilement enlevés.

Il existe trois modèles de grandeur normale, tous excités par un générateur à ultra-sons type L.325, donnant une sortie de puissance d'environ 150 W par impulsions.

EE 24 767 *pour plus amples renseignements*

Résumés des Principaux Articles

Un système numérique de concentration de données à action rapide par H. Fuchs et J. R. Wastell

Résumé de l'article aux pages 468 à 472

Cet article décrit un système de concentration de données, à action rapide, qui a été étudié pour contrôler les sorties d'un grand nombre de transducteurs électriques tels que couples thermo-électriques, extensomètres, thermomètres à boule de résistance, transducteurs de pression, etc.

Les modules sur lesquels ce système est fondé sont examinés et leur rôle est discuté.

L'article donne, en particulier, un aperçu du dessin d'un appareil de balayage pouvant déclencher des signaux microvolts à d'assez grandes vitesses, du dessin d'un amplificateur de données convenant au fonctionnement discontinu et à niveau de bruit et de dérive extrêmement bas et, enfin, d'un convertisseur à grande vitesse d'éléments analogiques en éléments arithmétiques, avec une précision de 0,1 % et une sortie décimale.

Le système, qui a été prévu pour un fonctionnement de 24 heures, englobe ses propes dispositifs de contrôle automatique et il peut détecter des anomalies de marche et produire un rapport complet sur les conditions de fonctionnement à intervalles réguliers de temps.

Un générateur de bruits de basse fréquence par N. T. Slater

Résumé de l'article aux pages 473 à 475

Au moyen d'une technique d'échantillonnage, 90% de la puissance de bruit se trouvant dans une bande haute fréquence est convertie en bruit de basse fréquence. Cette technique permet une sortie de basse fréquence de niveau élevé sans utiliser un amplificateur c.c. à grand gain et sa caractéristique de dérive inhérente. Une commande simple et efficace maintient la sortie de source constante et une méthode est décrite pour la mesure précise de la puissance de sortie finale. La distribution d'amplitude de la sortie est égale à celle de la source de bruit primaire.

Le générateur fonctionne entièrement par des moyens électroniques, n'employant aucune pièce mécanique mobile.

Systèmes à transistors et transducteurs par D. A. Ramsay

Résumé de l'article aux pages 476 à 479

Les transistors sont d'un niveau limité de puissance et de tension, et ils sont sensibles aux variations de température. Les transducteurs, par contre, n'ont pas ces limitations particulières et, pour certaines applications, ces deux semi-conducteurs ont des propriétés mutuellement complémentaires, de sorte qu'une combinaison des deux offre des avantages par rapport à l'utilisation indépendante de l'un ou de l'autre. En particulier, un élément de commande à transistors peut être doté de transducteurs pour donner des amplificateurs qui ne sont pas limités en puissance ou en tension de sortie. En outre, des amplificateurs de puissance hautement efficaces peuvent être formés avec des circuits analogues à ceux des amplificateurs magnétiques, mais dans lesquels les éléments à transducteurs de sortie sont remplacés par des transistors de commutation ou par des redresseurs commandés au silicium. Dans ces circuits, les transducteurs peuvent convenir parfaitement comme éléments de commande. En ce qui concerne tout particulièrement les transistors de commutation, les propriétés utiles des transducteurs peuvent être exploitées pour réaliser de petits systèmes robustes et hautement efficaces.

Un détecteur de soufflures à courants de Foucault par E. Ruff

Résumé de l'article aux pages 480 à 483

L'examen des trous de boulons de cellules d'avions pour le repérage des soufflures exige une méthode non-destructive rapide. Cet article décrit un instrument de ce genre, utilisant l'effet d'une spire court-circuitée brisée sur une inductance, ainsi que d'autres applications connexes.

Un circuit de commutation à transistors pour applications de régulation de puissance par M. J. Wright

Résumé de l'article aux pages 484 à 487

Un circuit amplificateur à transistors à courant continu, pour des applications de commande de puissance, est décrit dans cet article. L'oscillation continue de l'amplificateur est obtenue par réaction positive, l'étage de sortie étant commuté entre l'état de saturation et celui de coupure. Le signal d'entrée module le rapport marque/espace et fait ainsi varier le courant moyen injecté à une charge amplificatrice inductive. L'application première envisagée pour ce circuit est la régulation de tension de dynamos ou d'alternateurs en parallèle, mais une brève mention est faite, cependant, de son utilisation pour la régulation d'alimentations par batterie ou secteur.

Les principaux avantages d'un amplificateur de commutation sont la grande efficacité et la faible dissipation de puissance dans le transistor à commande série.

Démodulation sensible aux phases, avec interpolation par J. S. Johnston

Résumé de l'article aux pages 488 à 490

Cet article décrit une méthode pour obtenir une approximation beaucoup meilleure de la forme d'onde d'enveloppe d'une onde porteuse basse-fréquence modulée, qu'on ne pourrait le faire par la démodulation sensible aux phases. Cette méthode s'appuie sur la connaissance préalable de la forme nominale de la forme d'onde, et une attention particulière est apportée au cas sinusoïdal, fort important.

Une source simple de courant constant à alimentation compensée par P. Pargas

Résumé de l'article aux pages 490 à 492

L'auteur décrit ici un circuit simple fournissant un courant d'une très haute stabilité pour le curseur de mesure d'un potentiomètre auto-équilibré. La stabilité totale à long terme, obtenue durant une période de 1000 heures, est supérieure à 0,15%, malgré les fluctuations de secteur et les variations de température.

La correction des erreurs dans les générateurs de fonctions à potentiomètre par J. Merchant

Résumé de l'article aux pages 493 à 496

Cet article décrit une méthode pour corriger les différentes erreurs qui peuvent se produire dans les générateurs de fonctions à potentiomètre, par suite du chargement du bras et d'autres restrictions d'ordre pratique.

Les erreurs causées par approximation d'une fonction continue avec des segments de ligne ne sont pas considérées.

Dans cette méthode, le potentiomètre est chargé de résistances, étudiées de la manière la plus simple pour donner la fonction voulue de segment de ligne. La fonction résultante est mesurée et ses déviations de l'objectif du dessin sont utilisées pour calculer une série de conductances de correction pour les segments de potentiomètre.

Le potentiomètre, modifié par ces conductances, donnera alors une fonction beaucoup plus proche de l'objectif du dessin. Une seconde application de la méthode peut être faite pour améliorer encore la précision de la fonction, mais dans la plupart des cas un seul processus de correction suffira.

La production de transformations en séries et de coefficients de Fourier dans un calculateur analogique par F. C. Harbert

Résumé de l'article aux pages 496 à 498

Cet article décrit deux montages de calculateurs pour obtenir la transformation en séries de Fourier d'une fonction de temps. L'un de ces montages est à amplificateurs et multiplicateurs de calcul, tandis que l'autre n'utilise que des amplificateurs de calcul.

Cette entrée en matière est suivie par la description d'une méthode pour obtenir, dans un calculateur analogique, la réponse de fréquence d'un système physique, une fois connue sa réponse à une entrée pas-à-pas. Cette méthode est éclairée par un exemple. Une extension des montages de calcul permet d'extraire les coefficients de Fourier d'une forme d'onde périodique.

Un amplificateur intégrateur et différenciateur pour les mesures de variations par M. Kringlebotn

Résumé de l'article aux pages 504 à 505

La source de sigaux dans les mesures de vibrations est tenue comme étant un capteur de vitesse, et la puissance de sortie de l'amplificateur comme étant débitée à une charge à basse impédance (comme, par exemple, un galvanomètre à résistance interne de 25 ohms). Lorsque l'amplificateur est employé comme intégrateur, il convertit le signal de vitesse d'entrée en un signal indiquant le déplacement, et lorsqu'il est utilisé comme différenciateur, il convertit le signal de vitesse d'entrée en un signal indiquant l'accélération du capteur. L'amplificateur peut aussi être utilisé pour l'amplification linéaire du signal de vitesse.

Les étages amplificateurs sont stabilisés par rapport aux changements de température. La réaction supplémentaire contribue pareillement à la stabilité thermale. Une source unique de 12,6V est utilisée.

Un nouvel oscillateur résistance-capacité à transistor unique par E. T. Emms

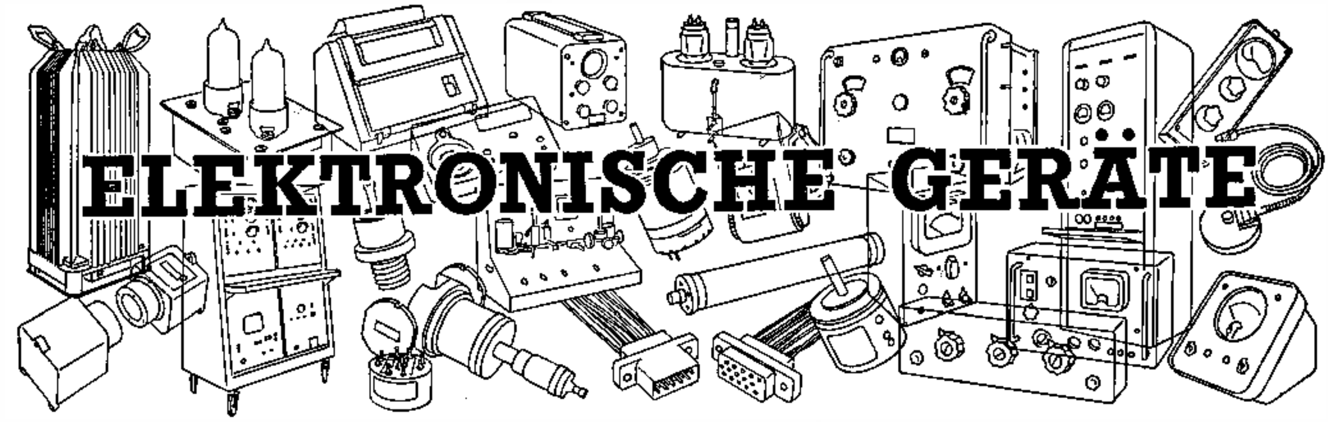
Résumé de l'article aux pages 506 à 508

L'auteur décrit un nouvel oscillateur résistance-capacité à transistor unique, qui oscille avec la plupart des transistors à basse fréquence, même lorsque le gain de courant du transistor est à la limite inférieure. Le circuit emploie un réseau à parallèle T modifié. L'auteur discute en conclusion l'effet de l'impédance d'entrée de transistor finie et de la capacité de sortie.

L'utilisation d'un transistor pour la commutation d'un noyau magnétique à antenne cadre par I. Krajewski

Résumé de l'article aux pages 509 à 511

Après avoir montré que la plupart des transistors pouvant être obtenus doivent être considérés comme des sources de tension lors de la commutation de noyaux, l'auteur analyse le circuit de base pour régler un noyau. Cette analyse démontre que, dans des conditions déterminées et pour réaliser une commutation très rapide, il existe un nombre de tours optimum pour le bobinage de commutation. Une vérification expérimentale de ces résultats, qu'indique l'auteur, concorde fort bien avec les calculs. Une brève description d'un circuit pratique est donnée en conclusion, ainsi que les calculs de dissipations de puissance de jonction à l'intérieur du transistor.



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 516 bis 519

ANALOGRECHNER FÜR VERFAHRENSREGELUNG

De Havilland Propellers Ltd. Hatfield Aerodrome, Hatfield, Hertfordshire

(Abbildung Seite 516)

Es wird gesagt, dass „Anatrol" eine neue Entwicklungsrichtung im Analogrechnerbau darstellt und zum ersten Mal einen preiswerten, umfassenden und vielseitigen Rechner zur Verfügung stellt, der die für Verfahrensregelung unerlässliche Stabilität und stossfeste Ausführung hat.

Die Rechengenauigkeit des „Anatrol" entspricht der modernen Messgenauigkeit und ermöglicht seine bequeme und wirksame Integrierung in bestehende Regelungssysteme.

Die Geschwindigkeit, mit der übliche Analogrechner Verfahrensgleichungen auflösen, ist im Vergleich mit der Verfahrensdynamik äusserst schnell. Diese Tatsache führte zusammen mit dem Bedürfnis, die Betriebssicherheit zu erhöhen und den Preis herabzusetzen, zur Wahl der verwandten Grundzüge.

Eine arithmetische Einheit wird aus einem Minimum von Rechenverstärkern und Multiplizierern gebildet. Diese Einheit wird dann dazu benutzt, die Gleichungen stufenweise nach einem vorgewählten Programm aufzulösen, in dem jede Stufe durch einen Folgeschalter eingestellt wird. Kleine Speicher werden zum Festhalten des Resultats einer Stufe der Lösung eingesetzt und diese Daten in einem späteren Stadium der Rechnung verwandt. Grössere Speicher werden zur Ausgabefestlegung der letzt berechneten Zahl benutzt, während neue Werte bestimmt werden.

Bis zu 25 Stufen stehen zur Verfügung, was dem Rechner Verwendungsmöglichkeiten gibt, die einer Anlage mit mindestens 75 Rechenverstärkern und 25 Multiplizierern entspricht. Mit Ausnahme der elektromechanischen Schaltelemente wird durchgehend Halbleitertechnik benutzt. Der Rechner ist nach dem Bausteinprinzip konstruiert, und je nach Problem können verschiedene Bauelementkombinationen eingesetzt werden. Auf Wunsch sind zusätzliche Ein- und Ausgabegeräte mit Wandlern und Verstärkern lieferbar, die den Rechner genau an die Mess- und Reglergeräte im Betrieb anpassen.

Der Rechner hat eingebaute Eigenprüfschaltungen. Für Fehler über 1% wird eine Warnung gegeben und der Rechner automatisch vom Verfahren isoliert.

EE 24 751 *für weitere Einzelheiten*

SCHWINGUNGSTOLERANZ-MONITOR

Dawe Instruments Ltd, 99 Uxbridge Road, London W.5

(Abbildung Seite 516)

Der Schwingungstoleranz-Monitor Type 1434 wurde zur Messung, Überwachung und Aufzeichnung von Schwingungen von Maschinen und Bauten eingeführt. Sowohl grosse Rotationsmaschinen, einschliesslich hydro- und turboelektrische Anlagen, als auch schwer belastete Brücken und Dämme können automatisch auf übermässige Schwingungen überwacht werden, indem das vom Geber abgegebene Signal laufend mit einer eingestellten Vorspannung verglichen wird. Eine Auslöseröhre, in deren Anodenkreis ein Leistungsrelais liegt, wird getriggert, sobald der Signalpegel über einen vorbestimmten Wert ansteigt. Das Relais schaltet entweder die Maschine ab oder gibt Warnung.

Der Schwingungstastmesser hat eine robuste Tauchspulenkonstruktion und ist hermetisch verschlossen, um bei hoher Temperatur und Feuchtigkeit die Genauigkeit aufrecht zu erhalten. Wenn der Tastmesser in gefährlicher oder unzugänglicher Lage an Maschinen angebaut ist, kann der Monitor mit Hilfe zweiadriger Schirmleitungen bis zu 60 m entfernt in einer sicheren Position aufgestellt werden. Über einen Frequenzbereich von 20 ... 400 Hz (1200 ... 24 000 UPM) gibt der Tastmesser eine Spannung ab, die der Schwingungsgeschwindigkeit der Oberfläche, mit der er in Berührung steht, proportional ist. Wenn die Ausgangsspannung an ein summierendes Netzwerk angelegt, dann verstärkt und gleichgerichtet wird, erhält man ein Signal, das die Verschiebung darstellt. Dieses Gleichstromsignal wird auf einem Messgerät angezeigt, das in Verschiebungsspitzenwerten einer auf den Tastmesser einwirkenden einfachen harmonischen Bewegung geeicht ist. Anzeige des Durchschnitts-Schwingungsniveaus kann durch Einstöpseln eines normalen registrierenden Gleichstrom-Milliammeters aufgezeichnet werden.

Ein Empfindlichkeitsregler stellt vier Bereiche ein: 0 ... 0,001 Zoll; 0 ... 0,003 Zoll; 0 ... 0,010 Zoll und 0... 0,030 Zoll (englische Masse) oder 0 ... 0,03 mm; 0 ... 0,1 mm; 0 ... 0,3 mm und 0 ... 1,0 mm (metrische Masse). Der Monitor kann bei jedem Schwingungsniveau zwischen 20 ... 100% Vollausschlag für Triggerung eingestellt werden.

EE 24 752 *für weitere Einzelheiten*

FREQUENZANALYSATOR

Laboratoire Electro-Acoustique, 5, rue Jules-Parent, Rueil (Seine-et-Oise), Frankreich

(Abbildung Seite 516)

Das Gerät wurde zur Analyse von Komponenten eines Frequenzspektrums mit einer Grundfrequenz von 50 Hz ausgelegt.

Die Analyse wird auf folgende Weise durchgeführt:

Ein im Analysator eingebauter Präzisionsgenerator erzeugt eine veränderliche Frequenz, die dem zu analysierenden Spektrum überlagert wird. Nach Gleichrichtung wird das Signal einem mit Messgerät ausgerüsteten selektiven Verstärker zugeführt, der nur ein enges Band durchlässt. Wenn die Differenz zwischen der veränderlichen Frequenz und der Spektrumkomponente eine Frequenz im Durchlassbereich ergibt, wird das Instrument proportional zur Komponentenamplitude abgelenkt.

Mit dem Analysator F.A.H.17 können entweder die Komponenten nacheinander oder eine einzelne Komponente in Bezug auf Zeit untersucht werden.

Da die 50 Hz Grundfrequenz nicht ganz konstant sein könnte, ist es nötig, ein Durchlassbereich gewisser Breite vorzusehen, so dass die verschiedenen Komponenten abgestimmt werden können. Mit zu scharfer Abstimmung können selbst sehr kleine Frequenz-schwankungen Schwankungen der Instrumentanzeige hervorrufen, ohne dass tatsächlich eine Änderung der Komponentenspannung vorliegt.

Die Trennschärfe des Gerätes, mit Ausnahme des oben erwähnten Durchlassbereiches, muss jedoch genügend hoch sein, um die Beeinflussung der zu untersuchenden Komponente durch beiderseitig benachbarte Komponenten zu verhindern.

Der Frequenzbereich ist 100 Hz... 2 kHz und die Dämpfung 25 dB bei ±50 Hz der Mittelfrequenz des Durchlassbereiches. Anschlussklemmen für den Betrieb eines Schreibers sind vorgesehen.

EE 24 753 *für weitere Einzelheiten*

KLEINST-POTENTIOMETER

Miniature Electronic Components Ltd, The Lye, St. Johns, Woking, Surrey

(*Abbildung Seite 517*)

M.E.C. Ltd stellen unter Lizenz von Con-Elco in Kalifornien ein Programm von Kleinst-Einstellpotentiometern her, die in drei Ausführungen lieferbar sind.

Der Temperaturbereich liegt zwischen −55°C...+180°C, Nennbelastung ist 2¼ W und geht bei 150°C auf ½ W herunter. Der Widerstandsbereich geht von 10 Ω ... 50 kΩ. Diese Bauelemente wurden besonders für die Fernwaffen- und Raketenindustrie entwickelt, wo schärfste Klimabedingungen bestehen und äusserste Betriebssicherheit von grösster Wichtigkeit ist.

EE 24 754 *für weitere Einzelheiten*

NF-LEISTUNGSMESSER

Amos of Exeter Ltd, Weircliffe Court, Exwick, Exeter, Devonshire

(*Abbildung Seite 517*)

Die Konstruktion des Modells 157 lehnt sich eng an die der anderen WEIRCLIFFE NF-Leistungsmesser an. Die Skala des 114 mm Drehspulinstruments ist für 0...10 mW und −20 dB...+10 dB geeicht. Alle Belastungswiderstände sind induktionsfrei gewickelt, für Dauerbelastung bemessen und mit einer Genauigkeit von 0,1% abgeglichen. Einzelabweichungen des Messwerks können durch unabhängige Regelung ausgeglichen werden.

Die Eingangsimpedanz wird mittels eines grosszügig bemessenen C-Kern-Übertragers angepasst, dessen Kernbandstärke von 0,1 mm einen linearen Frequenzgang ergibt. Für einen vollsymmetrischen Eingang für alle Impedanzen ist eine Mittelanzapfung vorgesehen. Die Isolation ist so bemessen, dass das Gerät direkt an eine Röhrenschaltung angeschlossen werden kann, um die besten Belastungsbedingungen zu bestimmen. Zwei Impedanzschalter stellen gewählte Werte zwischen 1 Ω ... 10 Ω in Stufen von 0,5 Ω; Multiplikatoren für ×1, ×2, ×5, ×10, ×20, ×50, ×100, ×200, ×500 und ×1000 ermöglichen schnelle Einstellung jeder Impedanz von 1 Ω ... 10 kΩ.

Ein fünfstufiger Schalter stellt Bereiche für 10 mW, 100 mW, 1 W, 10 W und 50 W Vollausschlag ein. Für jeden Bereich ist die niedrigste brauchbare Anzeige ein Zehntel des Vollausschlags.

EE 24 755 *für weitere Einzelheiten*

ZWÖLF-KANAL-SCHREIBER

AB Elektrisk Malmletning, Danderydsgatan 11, Stockholm

(*Abbildung Seite 517*)

Die ABEM-Serie direktschreibender Registriergeräte besteht aus Präzisionsinstrumenten für direkte Aufzeichnung von zwischen 1...12 Vorgängen niedriger Geschwindigkeit; diese Vorgänge können dem mit Mittelanzapfung versehenen Drehspulgalvanometer als Gleich- oder Wechselstromsignale zugeführt werden. Die direkte Aufzeichnung wird durch beheizte Schreibstifte auf wachsbeschichtetem Papier erzielt. Die Wachsschicht schmilzt, wo sie beim Durchlaufen mit den Schreibstiften in Berührung kommt, und das glänzende rote Papier wird sichtbar und bildet eine stark abstechende Spur. Zwei vorgesehene Markiereinrichtungen werden zum Aufzeichnen von Zeit- und Bezeichnungsmarken der Vorgänge benutzt.

Durch Wechselgetriebe kann jede der sich ergänzenden drei Geschwindigkeiten zwischen 1 mm/min und 10 cm/sec eingestellt werden. Galvanometer können mit einem Innenwiderstand von 1750 Ω, 430 Ω, 100 Ω oder 28 Ω mit Empfindlichkeiten von 25 mm/1 mA, 25 mm/2 mA, 25 mm/4 mA und 25 mm/8 A geliefert werden. Die Registrierpapierrollen sind 90 m lang und 383 mm breit. Die Schreibstifte haben eine Höchstablenkung von 25 mm nach beiden Seiten einer Nullstellung.

EE 24 756 *für weitere Einzelheiten*

REFLEXKLYSTRON-OSZILLATOR

English Electric Valve Co. Ltd, Chelmsford, Essex

(*Abbildung Seite 517*)

Das Niederspannungs-Reflexklystron K361 wurde besonders für Verwendung in Doppler-Geschwindigkeitsmessystemen entwickelt, in denen zuverlässige Arbeitsweise bei geringster Wartung von grösster Bedeutung ist.

Zu typischen Einsatzbeispielen gehören ein von der Westinghouse Brake and Signal Company entwickeltes Gerät für automatische Geschwindigkeitsregelung beim freien Rangieren von rollendem Material in Verschiebebahnhöfen und ein „Radar"-Doppler-Geschwindigkeitsmesser für die Polizei. Die mechanische Durchstimmung über den Frequenzbereich 10 700 MHz...10 725 MHz erfolgt mittels einer einzigen Schraubenabstimmung, die in die Kammer hineinragt.

Die Mindestleistung des K361 ist 20 mW. Es ist mit einem Kuppler Type UG-40 A/U (Z 830051) zur Verbindung mit Hohlleiter Nr. 16 ausgerüstet.

Für den Heizer werden 6,3 V 0,6 A benötigt, und übliche Betriebsbedingungen umfassen eine Schwingkammerspannung von 300 V, einen Schwingkammerstrom von 25 mA und eine Reflektorspannung von −200 V.

EE 24 757 *für weitere Einzelheiten*

STABILISIERTE NETZGERÄTE

International Electronics Ltd, 132-135 Sloane Street, London S.W.1

(*Abbildung Seite 517*)

Diese volltransistorisierten Bausteine sind auf 1,6 mm starken verzinkten Chassis aufgebaut, die entweder an eine Frontplatte oder auf eine Grundplatte montiert werden können.

Hergestellt werden z.Z. Bausteine für 100 mA, 200 mA, 350 mA und 500 mA Stromentnahme und 200 V, 250 V und 300 V Nennspannung. Ausserdem haben alle Geräte zwei 6,3 V 5 A Heizspannungswicklungen; eine derselben hat eine Mittelanzapfung.

Die Einheiten sind durch Netz- und Ausgangssicherungen geschützt.

Die Ausgangsimpedanz ist für 500 kHz unter 0,5Ω, die Effektiv-Restwelligkeit unter 1 mV und der Stabilitätsfaktor 400 : 1.

Die ersten Modelle einer transistorisierten Netzgerätserie für niedrige Spannungen sind auch lieferbar. Die Geräte geben für jede Spannung zwischen 5 ... 50 V 2 A ab und vermeiden durch eine Schalttechnik Verwendung vieler parallelliegender Transistoren. Erweiterung der Serie durch Geräte für Stromentnahme von 5 A und 10 A ist vorgesehen.

EE 24 758 *für weitere Einzelheiten*

LINEARES POTENTIOMETER

General Controls Ltd, 13-15 Bowlers Croft, Honywood Road, Basildon, Essex

(*Abbildung Seite 518*)

Dieses lineare Präzisions-Potentiometer mit entlastetem Aufhängeschleifer wurde entwickelt, um auch unter schwersten Stoss- und Schwingungsbedingungen die Schleiferabhebung vom Widerstandselement zu verhindern. In dem Instrument wird ein U-förmiges Schleiferelement zwischen der Kollektorstange und dem Widerstandselement im Druckgleichgewicht gehalten. Jede Fremdkraft, die auf Änderung des Schleiferkontaktdruckes hinwirkt, ruft eine gleichgrosse entgegenwirkende Reaktion auf der Kollektorseite des U-Schleifers hervor, wodurch konstanter Schleifenkontaktdruck aufrecht erhalten wird. In einem System dieser Art sind Kontaktrauschen und Ungleichförmigkeit des abzugreifenden Signals äusserst gering, und Beschädigung des Widerstandselementes als Resultat des Schleiferdruckes wird vermieden.

Bei einem Widerstandsbereich zwischen 100 Ω ... 10 kΩ sind bis zu fünf Anzapfungen einschliesslich Endverbindungen möglich.

EE 24 759 *für weitere Einzelheiten*

AUTOMATISCHES MW-PEILGERÄT

Marconi's Wireless Telegraph Co. Ltd, Chelmsford, Essex

(*Abbildung Seite 518*)

Das Peilgerät Type ND 103 benutzt das Bellini-Tosi System mit einem festen Rahmen und automatisch eingestellter Goniometerspule, die nach dem Nullverfahren arbeitet.

In einem einzigen kompakten Gehäuse und für Tischmontage geeignet, sind ein empfindlicher Empfänger und das Radio-Goniometer untergebracht. Das Gerät ist hauptsächlich für automatisches Arbeiten konstruiert, jedoch auch mit Handbedienung ausgerüstet. Die wichtigsten Regelelemente für automatischen Betrieb sind leicht zugänglich und unter einer Klappe bequem angeordnet.

Auf der für leichte Ablesung geneigten Frontplatte befindet sich links die Abstimmskala und rechts die Peilanzeige mit dem Abstimmeter zwischen den Skalen.

Das Gerät ist zum Einbau mit allen führenden Kreiselverstärkermodellen geeignet, und dadurch braucht der Steuerkurs des Schiffes nicht festgestellt zu werden. Es entspricht in jeder Beziehung den Vorschriften der englischen Postbehörden für Schiffsgeräte.

Für automatischen Betrieb ist die Peilgenauigkeit besser als ±1%, und ein eingebauter Quarzeicher dient der Skalenprüfung.

Unverzügliche Umschaltung von "automatisch" auf "manuell" ist vorgesehen, um schwache Signale in Gegenwart starker Signale, z.B. bei Notsignalen, anpeilen zu können.

Der Empfängerausgang speist einen Lautsprecher, Kopfhörerbuchsen sind jedoch auch vorgesehen. Der Frequenzbereich liegt zwischen 250 ... 550 kHz.

EE 24 760 *für weitere Einzelheiten*

P.V.C.-HOCHTEMPERATURSCHLÄUCHE

Permanoid Ltd, New Islington, Manchester, 4

Permanoid hat für Hochtemperaturverwendung oder -bearbeitung einen neuen P.V.C.-Schlauch mit Viskoseseidearmierung eingeführt, der der Britischen Norm BS 2848:1957, Type 1, Klasse 105T entspricht.

Der Permanoid Viskoseseideschlauch verbindet die Vorteile der eingebauten Textilieneinlage und des P.V.C.-Mantels, ohne irgendwelche der Nachteile zu haben. Viskoseseideverstärkung gibt Biegsamkeit in der Benutzung und verhindert Schrumpfen während des Ofenhärtens. Ein aus Weich-P.V.C. bestehender äusserer Überzug erhöht die Dichte und führt zu grösserem Widerstand gegen Tränköle oder -lacke, wodurch die Gefahr des mechanischen Versagens während der Bearbeitung effektiv verringert wird. Er hat gute Alterungseigenschaften, ist tropenschimmelfest usw. und für Dauerbetrieb in Temperaturen bis zu 105°C geeignet.

Er wird in elf Farben (einschliesslich schwarz und weiss) und für Innendurchmesser von 0,5 ... 5,0 mm geliefert.

Bei Prüfung nach BS 2848 ist der Isolationswiderstand nicht unter 1 000 MΩ und die Durchschlagsspannung 3 kV.

EE 24 761 *für weitere Einzelheiten*

HF-BRÜCKE

Hatfield Instruments Ltd, Crawley Road, Horsham, Sussex

(*Abbildung Seite 518*)

Die HF-Brücke ist direkt in fünf Bereichen für Widerstände von 10Ω ... 10 MΩ, Kapazitäten von 0 ... 0,025 μF (niedrigster Bereich 0...2pF), Leitwerte von 0...100 mmho (niedrigster Vollausschlag 0,01 mmho), Induktivitäten von 0,4 μ ... 100 mH und Blindwiderstände von 4 Ω...1 MΩ geeicht. Alle Messungen können über einen Frequenzbereich von 15 kHz ... 15 MHz vorgenommen werden; die Widerstands-, Kapazitäts- und Leitwertbereiche sind frequenzunabhängig; die Induktivitäts- und Blindwiderstandseichung bezieht sich auf $\omega = 10^7$.

Auf der schrägen Frontplatte sind zwei geeichte Skalen, die beiden Bereichschalter, Nulleinstellung und zwei koaxiale Buchsen für Quelle und Detektor angeordnet. Auf der oberen Platte befinden sich sechs Buchsen für Erd-, Null-, Gemeinschafts- und drei Bereichsverbindungen zu den zu messenden Geräten oder Bauelementen.

Verbindungen werden mittels gefederter Einschubstecker leicht hergestellt. Die Einschubmethode ist auch für schnelles Auswechseln von Vorrichtungen in Reihenprüfung sehr nützlich. Als Zusatzgerät lieferbare Transistor-Anpassungsgeräte können auch direkt eingeschoben werden.

Die Brücke kann mit jeder geeigneten Quelle und Anzeige benutzt werden. Für Betrieb mit fester Frequenz wurde für die Brücke der Quellen- und Anzeigebaustein LE 302 mit $\omega = 10^7$ entwickelt. Dadurch kann die Brücke direkt in Induktivität geeicht werden. Bei Einbau dieses Bausteins in die Brücke findet ein Milliammeter als Detektor Verwendung, das, wie in der Abbildung gezeigt, zwischen den beiden Skalen angeordnet ist.

EE 24 762 *für weitere Einzelheiten*

TRANSISTOR-TESTER

Microcell Ltd, Blackwater Industrial Estate, Camberley, Surrey

(*Abbildung Seite 519*)

Der Tester ist für Messung der Parameter von pnp und npn Transistoren mit hoher Genauigkeit konstruiert. Verschiedene im Gerät eingebaute Vorrichtungen werden besonders dem Schaltungsingenieur zustatten kommen, der genauere Information über einzelne Transistoren benötigt als die normalerweise vom Hersteller angegebenen. Er ist besonders zur Auswahl angepasster Transistor-Sätze geeignet.

Folgende Parameter können gemessen werden: Verstärkungsfaktor, Grenzfrequenz, Kollektorreststrom, Kollektor-Knickpunktspannung von Transistoren (Emitterschaltung) und Knickpunktspannung sowie Reststrom von Dioden.

Das Gerät ist durch Differentialeingang, Breitband-Röhrenvoltmeter als Verstärkungsanzeiger und einen Breitband-Oszillator als Quelle gekennzeichnet. Die Verwendung eines Röhrenvoltmeters als Anzeiger gleicht wirksam Fehler aus, die durch Störspannungen am Verhältnis-Netzwerk auftreten. Der Frequenzgang des Röhrenvoltmeters liegt ungefähr innerhalb 1 dB von 1 kHz...10 MHz. Der Oszillator bestreicht den Bereich von 1 kHz...10 MHz und gibt über diesen Bereich eine stabile Leistung innerhalb +0 und −1 dB ab. Leistungswechsel mit Änderung der Umgebungstemperatur werden durch eine Brücken-Reglerschaltung klein gehalten.

Völlig stabilisierte Stromversorgungen stehen für den Kollektor mit 2...10 V 3 A, 0...10 V und 0...100 V bei 30 mA sowie für Oszillator und Röhrenvoltmeter mit 220 mA zur Verfügung.

Zur Überwachung von Kollektorspannung und -strom sind Profilmessgeräte übereinander gestapelt, so dass beide gleichzeitig beobachtet werden können. Alle Messgeräte haben vollen Überlastungsschutz. Der Basisstrom kann auf externen Instrumenten überwacht werden.

EE 24 763 *für weitere Einzelheiten*

POTENTIOMETER-SCHREIBER

W. H. Joens & Co. GmbH, Martinstrasse 55, Düsseldorf

(*Abbildung Seite 519*)

Der Linienschreiber ist in einem Gehäuse 144 × 144 mm untergebracht und hat eine Schreibbreite von 100 mm. Das Gerät ist volltransistorisiert. Im Interesse von Betriebssicherheit und Wartung werden keine Elektrolytkondensatoren verwandt, und statt einer Normalbatterie wird eine temperaturkompensierte, transistorstabilisierte Spannungsquelle benutzt.

Der kleinste Messbereich beträgt 5 mV bzw. 10 μA Gleichstrom, und die Einstellgeschwindigkeit für Vollausschlag ist 1 s. Die Standardausführung ist mit Wechselgetrieben für acht Papiervorschübe zwischen 10 ... 1200 mm/h und einem Wähler für sechs verschiedene Spannungen zwischen 24 ... 250 V ausgerüstet. Ein zweiter Schleifdraht und ein einstellbares Sollwertpotentiometer sind eingebaut, um eine Ausgangsbrücke zur Anzeige der Sollwertabweichung zu bilden.

Verschiedene mit Kunstharz vergossene Steckeinheiten für Eingangsschaltungen und Verstärker sind nach Bedarf lieferbar.

EE 24 764 *für weitere Einzelheiten*

WERKZEUGMASCHINEN-STEUERUNG

The General Electric Co. Ltd, Magnet House, Kingsway, London W.C.2

Ein neues Geräteprogramm für Positions- und Folgesteuerung von Werkzeugmaschinen kann von der Electronics Division der General Electric Company Ltd bezogen werden.

Beide Systeme verwenden „Logik-Bausteine" und benutzen transistorisierte gedruckte Schaltungen und Kaltkathodenröhren. Sie sind bedeutend kleiner als herkömmliche Geräte, und das praktische Wegfallen elektromechanischer Steuervorrichtungen reduziert die Wartung auf regelmässige elektrische Prüfungen.

Positionsablesung erfolgt durch rotierende Digitalwandler, deren digitales Ausgangssignal die Arbeitstischeinstellung in zwei Richtungen angibt und die Leitspindelantriebsmotoren steuert. Die Steuerung erfolgt mit zwei Geschwindigkeiten: 229 cm/min für schnellen Durchlauf und einer langsameren Geschwindigkeit von 30,5 cm/min zur endgültigen Einstellung der gewünschten Position mit ±0,0125 mm Endgenauigkeit. Das Gerät wird bereits erfolgreich mit der Newall Lehrenbohrmaschine 2436 eingesetzt.

Mit der Folgesteuerungsanlage werden sowohl Zeitsteuerung wie auch Triggerung für mehrere gleichzeitige oder aufeinander folgende Arbeitsgänge ausgeführt.

Unter den ausgeführten Installationen befinden sich eine senkrechte Räummaschine und eine mittelgrosse Spritzgussmaschine.

EE 24 765 *für weitere Einzelheiten*

FELDAUSWERTEGERÄT

Servomex Controls Ltd, Crowborough, Sussex

(Abbildung Seite 519)

Dieses neue Widerstandsanalog ist eigentlich eine zweidimensionale Ausführung des elektrolytischen Tanks; mit Hilfe des Auswertegerätes kann eine Karte der Potential- und Flusslinien bequem und ohne besondere Erfahrung gezeichnet werden.

Bei Einsatz überlegt sich der Benutzer sein Problem und verwandelt die Mengen in elektrische Feldgrössen. Leitende Elektroden werden in jeder gewünschten Form mit Silberfarbe in einer Ebene auf Teledeltos-Papier aufgetragen. Die Elektroden werden mit dem FP92 verbunden und mit Hilfe eines genau geeichten Potentiometers Schleifdrahtanzeigen eingestellt. Die Anzapfung des Schleifdrahtes wird durch ein Symmetrie anzeigendes Galvanometer mit dem Messkopf verbunden. Der Benutzer tastet das Feld für Symmetrie ab, die mit einem Bleistiftpunkt am Messkopf gekennzeichnet wird. In wenigen Minuten ergeben sich genügend isoelektrische Linien für eine graphische und genaue Lösung des Problems. Im allgemeinen sind die Resultate auf 5% genau und mit besonderer Sorgfalt sogar besser.

EE 24 766 *für weitere Einzelheiten*

ULTRASCHALL-REINIGUNGSGERÄTE

Kerry's (Ultrasonics) Ltd, Warton Road, Stratford, London E.15

(Abbildung Seite 519)

Diese Junior-Serie von Ultraschall-Reinigungsgeräten eignet sich besonders zur Reinigung kleiner Teile wie z.B. Kugellager- und Kugelkäfigteile, Messwerke und -teile, Edelsteinlager, Uhren, Uhr- und Chronometerteile, Röhren- und Wasserreglerteile, Schmuckstücke während der Herstellung, optische Montagen einschliesslich Brillengestelle, Linsen usw.

Ausser für die oben genannten und viele anderen Elemente kann die Junior-Serie von Ultraschall-Reinigern auch im Labor zur Bewertung von Ultraschall-Techniken benutzt werden.

Die Betriebsfrequenz des Gerätes ist 40 kHz, und die Wandler sind in den Boden eines Behälters aus rostfreiem Stahl eingebaut. Der Behandlungsbehälter ist auf einen Sockel montiert, so dass das Gerät entweder auf einem Tisch aufgestellt oder in eine normale Uhrenreinigungsmaschine eingebaut werden kann, um einen Teil des normalen Reinigungstaktes zu bilden, der Waschen, Spülen und Trocknen einschliesst. In vielen Fällen können Teile ohne Demontage gereinigt werden, wodurch die Gefahr mechanischer Schäden vermieden wird. Klingende Schmutzfilme wie z.B. Polierrot, Tripoli usw. können leicht entfernt werden.

Drei Standardgrössen sind lieferbar, die alle vom Ultraschall-Generator Type L.325 getrieben werden und ungefähr 250 W Impulsleistung haben.

EE 24 767 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Schnelles Digital-Messwerterfassungssystem von H. Fuchs und J. R. Wastall

Das in diesem Beitrag beschriebene schnelle Digital-Messwerterfassungssystem wurde zur Überwachung der Messwerte zahlreicher elektrischer Geber wie Thermoelemente, Dehnungsstreifen, Widerstandsthermometer, Druckgeber usw. entwickelt.

Die Bausteine, aus denen das System aufgebaut ist, werden beschrieben und ihre Leistung besprochen.

Besonders eingehend behandelt wird die Konstruktion eines Abtastgerätes für ziemlich schnelles Schalten von Mikrovoltspannungen, eines Datenverstärkers für diskontinuierlichen Betrieb mit äusserst geringem Rauschen sowie Drift, und eines Analog-Digital-Wandlers mit 0,1% Genauigkeit und Dezimalausgabe.

Das System ist für 24-Stunden-Betrieb bestimmt und hat eingebaute Selbstkontrollen. Es kann Abweichungen von normalen Betriebsdaten entdecken und in regelmässigen Abständen eine vollständige Aufstellung der Daten herstellen.

Zusammenfassung des Beitrages auf Seite 468-472

Ein NF-Rauschgenerator von N. T. Slater

Mittels einer Abtasttechnik (sampling) werden 90% der in einem HF-Band vorhandenen Rauschleistung in NF-Rauschen umgewandelt. Das ermöglicht Verwendung eines Hochleistungs-Gleichspannungsverstärkers mit seinen Drifteigenschaften. Einfache und wirksame Regelung hält den Quellenausgang konstant, und ein Verfahren zur Messung der Endleistung wird beschrieben. Das Amplitudenspektrum ist dem der primären Rauschquelle gleich.

Der Generator arbeitet nur mit elektronischen Mitteln und hat keine beweglichen mechanischen Teile.

Zusammenfassung des Beitrages auf Seite 473-475

Transistoren und Transduktoren verwendende Systeme von D. A. Ramsay

Der Spannungs- und Leistungspegel der Transistoren ist begrenzt, und sie sind gegen Temperaturschwankungen empfindlich. Transduktoren sind nicht in dieser Weise beschränkt, und die beiden Elemente haben ergänzende Eigenschaften, so dass ihre Kombination Vorteile gegenüber Verwendung des einen oder anderen allein mit sich bringt. Das trifft besonders zu, wenn ein Transistor-Reglerelement mit Transduktoren eingebaut wird, wodurch Verstärker geschaffen werden, die bezüglich Ausgangsleistung oder -spannung keiner Einschränkung unterliegen. Auch können Hochleistungsverstärker aus Schaltungen aufgebaut werden, die denen magnetischer Verstärker analog sind, in denen jedoch die Ausgangstransduktorelemente durch Schalt-Transistoren oder gesteuerte Silizium-Gleichrichter ersetzt werden; in diesen Schaltungen sind Transduktoren als Reglerelemente sehr geeignet. Die nützlichen Eigenschaften der Transduktoren können besonders in Verbindung mit Schalt-Transistoren für kleine, robuste Systeme hoher Leistung verwertet werden.

Zusammenfassung des Beitrages auf Seite 476-479

Ein Wirbelstrom-Rissdetektor von E. Ruff

Zusammenfassung des Beitrages auf Seite 480-483

Untersuchung von Schraubenlöchern im Flugwerk auf Risse muss mittels schneller, nichtzerstörender Prüfung erfolgen. Der Beitrag beschreibt solch ein Gerät, das den Einfluss einer unterbrochenen Kurzschlusswindung auf eine Induktivität benutzt und gibt auch einige ähnliche Verwendungsmöglichkeiten.

Eine Transistor-Schleusenschaltung für Starkstromregelung von M. J. Wright

Zusammenfassung des Beitrages auf Seite 484-487

Eine transistorisierte Gleichstromverstärkerschaltung für Starkstromregelung wird beschrieben. Durch Mitkopplung wird der Verstärker zum ungedämpften Schwingen gebracht und die Endstufe zwischen übersteuertem und gesperrtem Zustand hin- und hergeschaltet. Das Eingangssignal moduliert das Impulsverhältnis und beeinflusst dadurch den Mittelwert des in eine induktive Verstärkerbelastung gespeisten Stroms. Die Schaltung ist in der Hauptsache für Spannungsregelung parallelgespeister Dynamos und Alternatoren gedacht; Verwendung für netz- oder batteriegespeiste, geregelte Kraftquellen wird kurz erwähnt.

Die Hauptvorteile des Schleusenverstärkers sind hohe Leistung und geringe Verlustleistung im seriengeschalteten Transistor.

Phasenempfindliche Demodulation mit Interpolation von J. S. Johnston

Zusammenfassung des Beitrages auf Seite 488-490

Der Beitrag beschreibt ein Verfahren, durch das eine wesentlich bessere Annäherung zur Hüllwellenform eines modulierten, niederfrequenten Trägers erzielt wird als mit üblicher phasenempfindlicher Demodulation. Das Verfahren setzt einige Vorkenntnis der nominellen Wellenform voraus, und eine wichtige sinusförmige Wellenform findet besondere Beachtung.

Eine einfache konstante Stromquelle mit Netzkompensation von P. Pargas

Zusammenfassung des Beitrages auf Seite 490-492

Die beschriebene einfache Schaltung gibt einen hochkonstanten Strom für den Messschleifdraht eines selbstabgleichenden Spannungsteilers ab. Die über eine Zeitspanne von 1000 Stunden erzielte Langzeitstabilität ist trotz Netz- und Temperaturschwankungen besser als 0,15%.

Fehlerberichtigung in Potentiometer-Funktionsgebern von J. Merchant

Zusammenfassung des Beitrages auf Seite 493-496

Eine Methode zur Berichtigung der verschiedenen Fehler wird beschrieben, die durch Armbelastung und andere praktische Beschränkungen auftreten können.

Fehler, die im geradlinigen Annäherungsverfahren für kontinuierliche Funktionen entstehen, werden nicht berücksichtigt.

Nach dem Verfahren wird das Potentiometer mit Widerständen belastet, die auf die einfachste Abschnittsfunktion Weise zur Erreichung der geradlinigen berechnet werden. Die entstehende Funktion wird gemessen und ihre Abweichungen vom Funktionsentwurf benutzt, um berichtigte Leitwerte für die Potentiometerabschnitte zu kalkulieren.

Das mittels dieser Leitwerte berichtigte Potentiometer ergibt nunmehr eine Funktion, die dem Entwurf viel genauer entspricht. Eine Wiederholung dieser Methode kann die Genauigkeit der Funktion weiter erhöhen; in vielen Fällen wird jedoch eine Berichtigung ausreichen.

Erzeugung von Fourierumwandlungen und -koeffizienen mittels eines Analogrechners von F. C. Harbert

Zusammenfassung des Beitrages auf Seite 496-498

Zwei Rechneranordnungen zur Darstellung von Fourierumwandlungen und -koeffizienten werden beschrieben. Eine Anordnung benutzt Rechnerverstärker und Multiplikatoren, die andere nur Rechnerverstärker. Eine Methode wird angegeben, nach der der Frequenzgang eines physikalischen Systems mittels eines Analogrechners dargestellt werden kann, sobald das Ansprechen auf eine Stufeneingabe bekannt ist. Diese Technik wird durch ein Beispiel erläutert. Eine Erweiterung der Rechneranordnungen erlaubt Errechnung der Fourierkoeffizienten einer periodischen Wellenform.

Ein summierender und differenzierender Verstärker für Schwingungsuntersuchungen von M. Kringlebotn

Zusammenfassung des Beitrages auf Seite 504-505

Es wird angenommen, dass ein Geschwindigkeitsgeber die Signalquelle für Schwingungsmessungen bildet und dass die Ausgangsleistung des Verstärkers einen Verbraucher niedriger Impedanz treibt (e.g. ein Galvanometer mit 25Ω Innenwiderstand). In dem zur Integration benutzten Verstärker wird das Geschwindigkeits-Eingangssignal in eine Anzeige des Weges umgewandelt; bei Einsatz für Differenzierung gibt das Signal eine Anzeige der Abtasterbeschleunigung. Der Verstärker kann auch für lineare Verstärkung des Geschwindigkeitssignals Verwendung finden.

Die Verstärkerstufen sind in Bezug auf Temperaturschwankungen stabilisiert. Zusätzliche Gegenkopplung trägt auch zur thermischen Stabilität bei. Eine einzige 12,6 V Spannungsquelle wird benutzt.

Ein neuartiger Einzeltransistor-RC-Oszillator von E. T. Emms

Zusammenfassung des Beitrages auf Seite 506-508

Ein neuartiger Einzeltransistor-RC-Oszillator wird beschrieben, der mit den meisten NF-Transistoren schwingt, selbst wenn die Stromverstärkung des Transistors an der unteren Toleranzgrenze liegt. Die Schaltung benutzt eine Abwandlung des Parallel-T-Netzwerkes. Der Einfluss von endlicher Transistor-Eingangsimpedanz und -Ausgangskapazität werden besprochen.

Verwendung eines Transistors zur Einstellung eines Magnetkernes mit rechteckiger Hysteresisschleife von I. Krajewski

Zusammenfassung des Beitrages auf Seite 509-511

Es wird gezeigt, dass die meisten lieferbaren Transistoren als Spannungsquellen betrachtet werden müssen, wenn sie Kerne einstellen. Danach wird eine Analyse der Grundschaltung für Kerneinstellung gegeben, die zeigt, dass unter den gegebenen Bedingungen für die Einstellwicklung eine optimale Windungszahl besteht. Die versuchsmässige Uberprüfung dieser Ergebnisse zeigt eine gute Ubereinstimmung mit der Kalkulation. Eine praktische Schaltung mit Berechnung der Verlustleistungen im verwandten Transistor wird beschrieben.