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Commentary

QUALITY and Reliability Year (QRY) was publicly announced some six months ago and is due to start next October; the campaign will have its official launching at the Festival Hall, London, on 20 October. It is being organized by the National Productivity Council, in association with the National Council for Quality and Reliability, in the firm conviction that here lies one of the most powerful weapons in British industry's fight against foreign competitors. In the words of HRH Prince Philip, "Manufacturers with a good reputation for quality and reliability have always found that this gives them a decided commercial advantage. At this particular moment when exporting and the earning of foreign exchange is so vital to the Nation's prosperity, it is essential that all British exporters should have a better reputation for quality and reliability than their foreign competitors.... A good reputation for well designed goods or components, fit for the purpose, which don't fail or break down, is the criterion for certain success."

Many firms already have well organized QR systems and departments and much has also been done in this direction by trades associations, research associations, professional bodies etc. The main objectives of the QRY will be to give the fullest publicity to the outstanding achievements of British firms in quality and reliability methods; to assist every firm that is not yet doing so to review its methods for achieving improved quality and reliability and the dissemination of knowledge to achieve these objectives. Although the year starting next October will see a sharp focus of attention on QR it is not, of course, intended that it should be a 'once for all' effort; for the campaign to be successful its results must be enduring.

The campaign has as its slogans 'right first time' and 'quality and reliability is everybody's business', and both are very apt. It is estimated that the manufacturing industry could make savings of between 1½ and 5 per cent of its gross turnover by making things right first time and avoiding waste and rework. Since the gross product of the British manufacturing industry is in excess of £10 000M per year, even at the lowest level this would represent a saving of some £150M p.a.; at the upper level the saving would be within the range of the visible trade gap. That 'QR is everybody's business' is demonstrably true. It is a matter of concern at all stages of a product from conception through design, manufacture and distribution and a matter which requires the genuine co-operation of workers at all levels, from the boardroom to the shopfloor. Indeed, in the final analysis QR depends almost entirely on people and one of the greatest tasks of management is to engender a real 'pride in the product' throughout the whole range of their activities.

The electronic industry is concerned with QR in two ways; externally and internally, so to speak. It can make a significant contribution towards quality, reliability and increased productivity in other industries by supplying measuring instruments and supervisory and control equipment ranging from a simple photo-electric relay to the largest computer. But, if it is to do this successfully, its own QR procedures must be beyond reproach. Indeed, so far as sales, either at home or abroad, of the more complex equipments are concerned quality, reliability and fitness for purpose are almost the only factors that count. Price, unless it is absurd, is of secondary, and one might almost say minor, importance. That is not to say that the lowest possible price should not be aimed at or that there is not a market for the cheaper types of instrument, but rather that a low price should not be the objective if it means any sacrifice in quality, reliability and most important of all, fitness for the purpose for which it is required or intended.

Fortunately the electronic industry, both in this country and overseas, is already more conscious of the necessity for effective QR procedures than is the case in many other industries. This, no doubt, is largely due to the fact that a great deal of electronic equipment is manufactured for military, and more latterly space, applications where reliability must not only be provided but must be specified in quantitative terms. In fact a number of manufacturers now regard the statistical estimation and specification of reliability, usually in terms of mean time between failures (m.t.b.f.), of both components and complete equipments, as a routine part of any design and development programme.

Of course, a great deal remains to be done and one of the most pressing needs is a standardized format and terminology for the presentation of reliability information. The British Standards Institution is not unaware of this problem and neither has it been idle. A technical committee on electronic reliability was formed and has been working on the subject since June 1964. In accordance with current BSI policy, and with export markets and international contracts in view, the committee has concentrated initially on co-operation with the corresponding technical committee of the International Electrotechnical Commission. Considerable progress has been made and work has now started on a British Standard the provisional title for which is "Basic Methods of Expressing Reliability Requirements for Electronic Equipment and the Components Used Therein ". It is hoped to circulate the first drafts for general comment before the end of this year.

The rewards of quality and reliability to user, manufacturer and to the country are plain to see but the attainments of high standards of QR are not easily come by and constant vigilance is needed to maintain them. QRY should do much to stimulate interest and endeavour in this direction and it would be wise for as many firms as possible to avail themselves of the services which are being made available by the National Productivity Council and the National Council for Quality and Reliability.

The Electronic Principles of the Mekometer II

By R. H. Bradsell*

The Mekometer II is an electronic distance measuring instrument designed to measure ranges up to 5000ft (1·5km) in the simplest possible manner.

It employs a light ray modulated at nearly 500Mc/s. Specially designed cavity resonators determine the modulation wavelength independent of a knowledge of the velocity of light and thus of the refractive index of the air along the ray path.

(Voir page 344 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 351)

THE Mekometer II was designed at the National Physical Laboratory to provide the surveyor, engineer, and architect with a portable electronic instrument which could measure distances up to 5000ft (1·5km), in the simplest possible manner. The instrument can obtain the required distance without any calculations or refractive index observations. Once set, the calibration of the device should remain constant for an indefinite period.

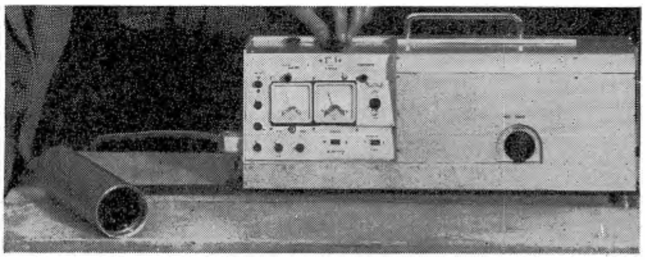


Fig. 1. The complete instrument

An accuracy of about 3 parts in a million has been obtained, with a sensitivity of better than 0·003ft (1mm).

The Mekometer II employs a light ray modulated at a frequency near to 500Mc/s. The modulation wavelength is held constant by specially designed cavity resonators independent of the refractive index changes along the ray path.

The instrument together with its power supply and the retro-reflector, which would normally be situated at the far end of the range to be measured, are shown in Fig. 1.

The Basic Method

A light beam whose polarization is modulated at u.h.f. is projected to a distant reflector whence it is returned

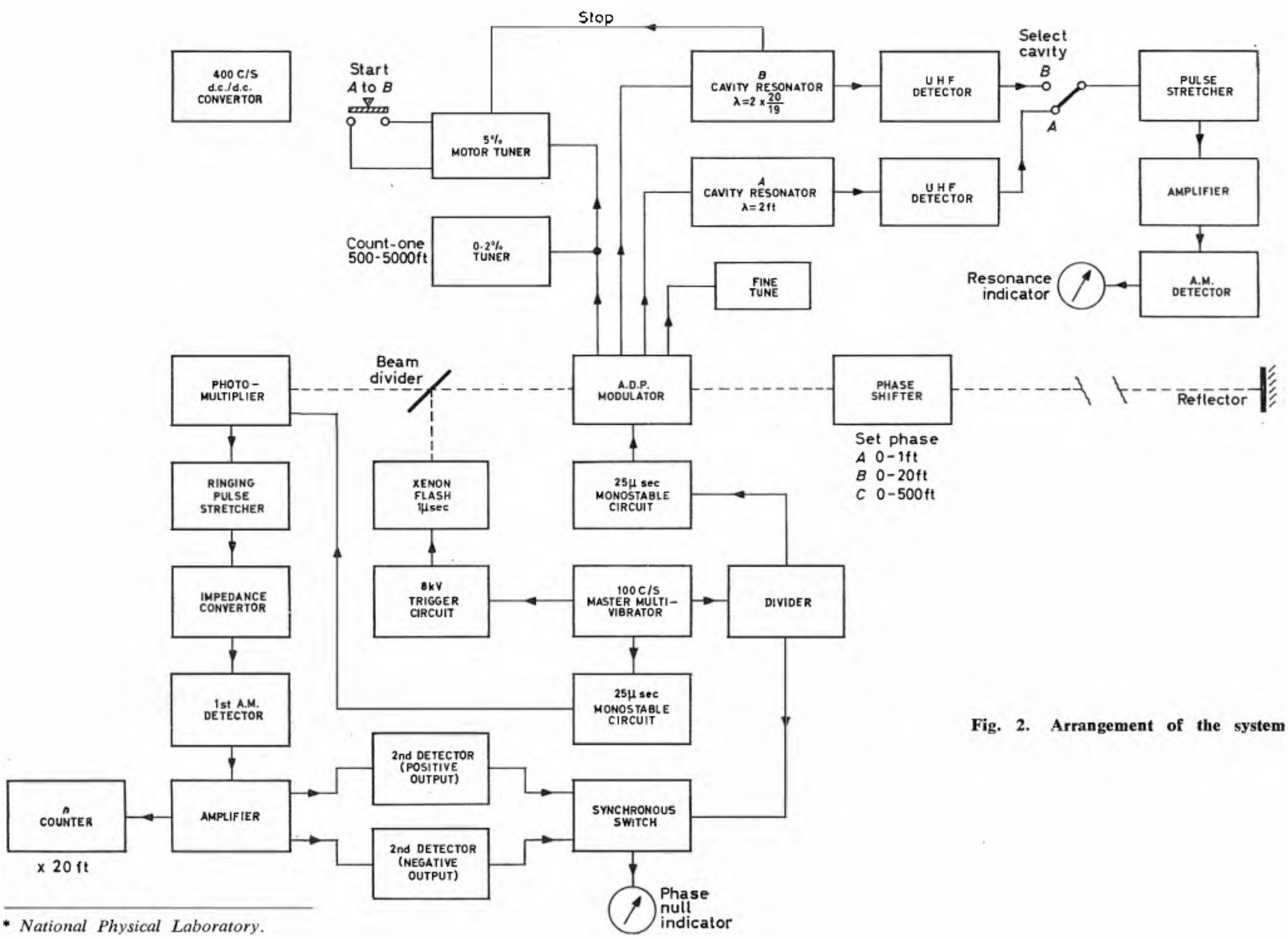


Fig. 2. Arrangement of the system

* *National Physical Laboratory.*

to the transmitter. The phases of the transmitted and received modulation are compared. If the ray path is varied mechanically until a phase balance is obtained and the modulation half-wavelength $\lambda/2$ is made for instance exactly 1ft, then the amount the path is varied measures the fractional part the distance is above an exact number of feet.

The modulation wavelength λ is monitored by quarter wavelength coaxial cavity resonators which are filled with dry air. These cavity resonators are allowed to acquire the atmospheric temperature and pressure of the ray path.

The modulation wavelength λ can be expressed as,

$$\lambda = f(L)\ \mu_r/\mu_g$$

where $f(L)$ is a linear function of the cavity dimensions and μ_r is the u.h.f. refractive index of the dry air in the cavities under the prevailing conditions. The group refractive index μ_g is the average value along the ray path for the light which is the carrier wave.

The ratio of the two refractive indices is substantially constant over a very wide range of conditions.

The Instrument

In the realization of the above principles into a viable electronic distance measuring instrument, every attempt was made to maintain the basic simplicity of the concept.

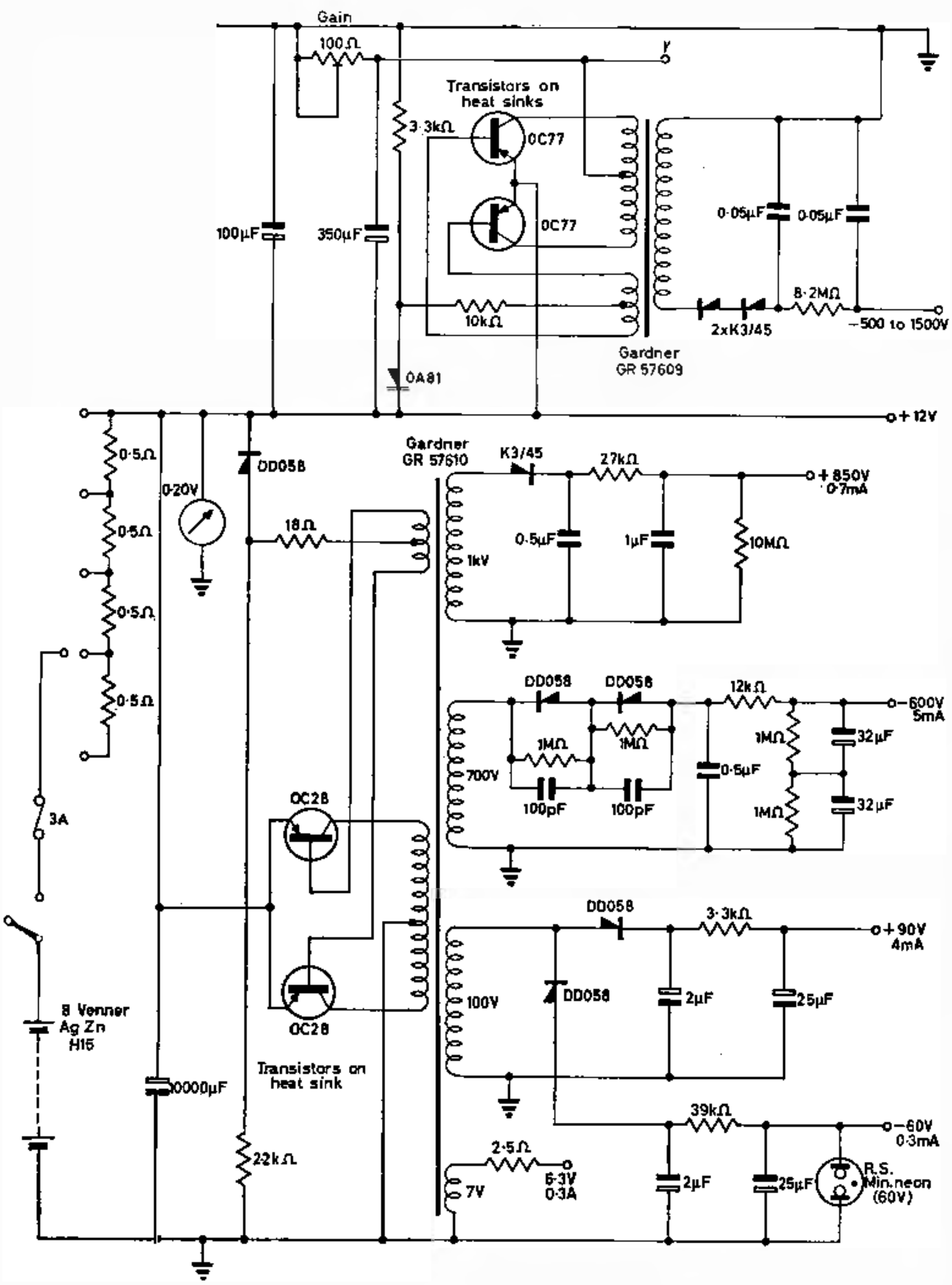
The heart of the Mekometer II is the a.d.p. (ammonium dihydrogen phosphate) light modulator which imposes elliptical polarization modulation on the transmitted light beam and demodulates it on its return from the distant target reflector. The arrangement of the electronic systems is shown in Fig. 2.

The modulator, which is a coaxial quarter-wave cavity resonator, is excited into strong u.h.f. oscillation by a pulsed triode valve. Under pulsed conditions the relatively low anode impedance allows matching to be achieved with the anode connected to the inner line near the short-circuit end of the cavity. A low Q tuned loop dips into the magnetic field inside the cavity and feeds r.f. energy to the valve cathode in such a phase as to enforce oscillation. Across the high voltage end of the coaxial resonator is situated the a.d.p. crystal which is the light modulator.

Light from the pulse xenon flash tube is rendered parallel and plane polarized. The beam-splitting reflector, with an aperture at its centre, projects an annular ring of light through the a.d.p. crystal where the u.h.f. electric field modulates it. A microscope objective forms a diminished real image of the source which in turn acts as a point source at the focus of a 2·5in telescope objective.

The projected light beam is directed towards the distant target reflector by way of a folded variable light path phase shifter. This light path

Fig. 3. The power supply



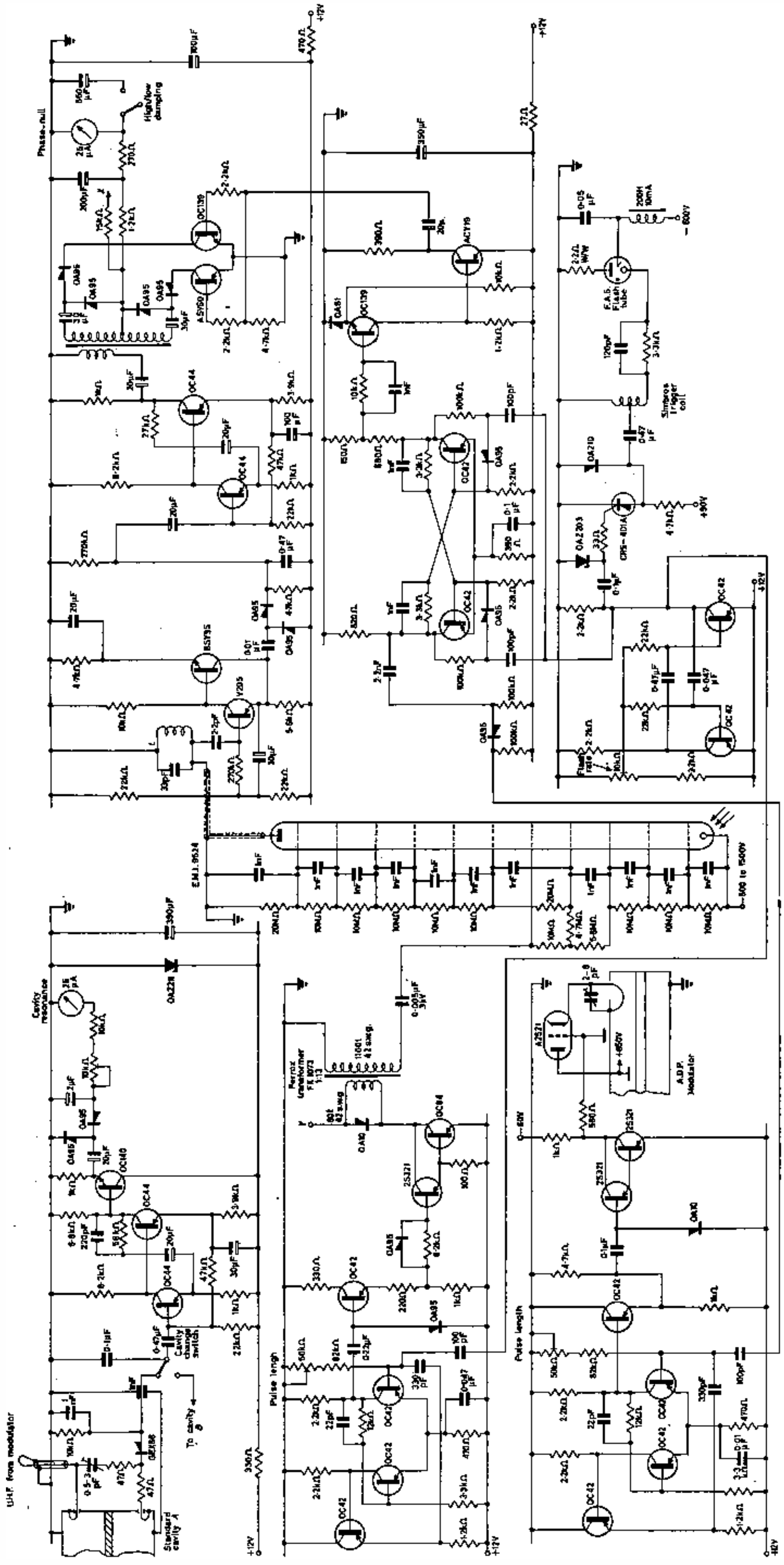


Fig. 4. The main circuits

allows the distance between modulator and reflector to be adjusted.

Returning light from the target reflector follows the same path and retraverses the crystal where the modulation is reinforced or diminished depending on the instantaneous phase of the electric field in the crystal relative to the phase of the modulation already imposed on the light. The beam is now transmitted through the aperture in the beam-splitting reflector and passes through a second, but crossed, polarizer. The intensity of the light now depends on the degree of elliptical polarization on the light. This intensity is detected by the photo-multiplier whose amplified output current is displayed on the phase-null meter.

For a given modulation wavelength the variable light-path phase shifter can be adjusted until the meter indicates a minimum intensity. The light-path shift is a direct measure of the excess of the distance over an odd number of modulation half-waves.

The absolute wavelength generated by the modulator is monitored by two standard coaxial cavity resonators designated *A* and *B*. These cavities have wavelengths of 2ft and $2 \times 20/19$ft exactly.

Advantage is taken of the fact that the air-spaced capacitor, formed between the end of the inner rod and the end plate of the outer sheath, increases the wavelength produced above that which would have been expected from the length of the inner line of the quarter-wave coaxial cavity. Now the inner lines are made of copper plated invar (low expansion alloy) and the outer sheaths of copper. The relative expansion between these two metals has the effect of reducing the end capacitance as temperature increases. If the end capacitance has the right absolute value, the capacitance change will exactly

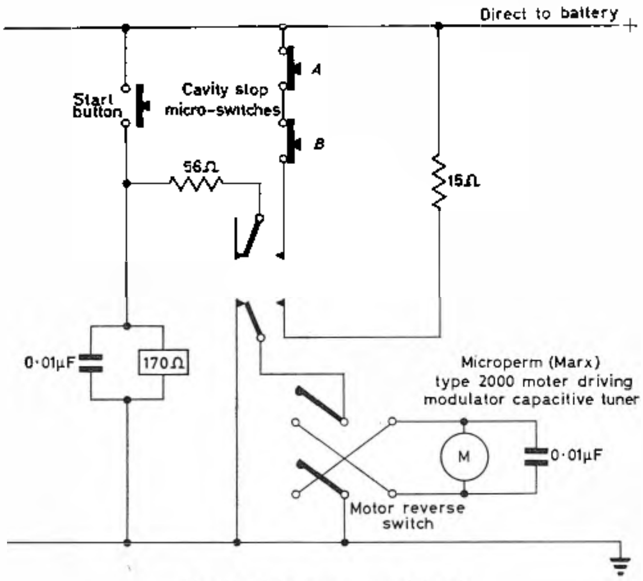


Fig. 5. The motor control circuit

compensate for the thermal expansion of the invar rod. The process is analogous to that of the compensated clock pendulum. Thus constant wavelength can be maintained as if the inner rod of the coaxial resonator had zero thermal expansion.

The air in the standard cavities is allowed to pick up atmospheric temperature easily by the situation of the cavities in the instrument. Atmospheric pressure is communicated by means of a soft plastic reservoir whose volume changes with the external pressure. The internal air is maintained dry by a connected container of silica gel.

A standard cats-eye target reflector is employed with the Mekometer II which has a 2·5in aperture $f/4$ lens. At the longer ranges a cluster of three such target reflectors is used. These reflectors are non-critical as to their alignment relative to the transmitted light beam.

The Mekometer II is provided with an eyepiece which, together with the main objective, form a telescope with a sufficiently wide field of view to locate the target reflector. When this has been done the returned light signal can be viewed, by means of a drop-in reflector, after it has retraversed the a.d.p. modulator so that the returned light spot can be centred in the aperture through the beam-splitting reflector before a measurement is commenced.

The Electronic Circuits

The power requirements of the instrument are supplied by a 12V, 15Ah silver-zinc accumulator. The basic 12V is converted to the various necessary potentials, see Fig. 3, by a transistorized d.c.-d.c. convertor operating at 400c/s. A second convertor supplies the high voltage needed for the photomultiplier and has a potentiometer controlled output, variable between 500 and 1 500V, which acts as a detector gain control. The total power consumption is about 12W, allowing up to 100 measurements per battery charge.

The main circuit arrangement is shown in Fig. 4. Apart from the u.h.f. planar triode modulator semiconductors are used throughout.

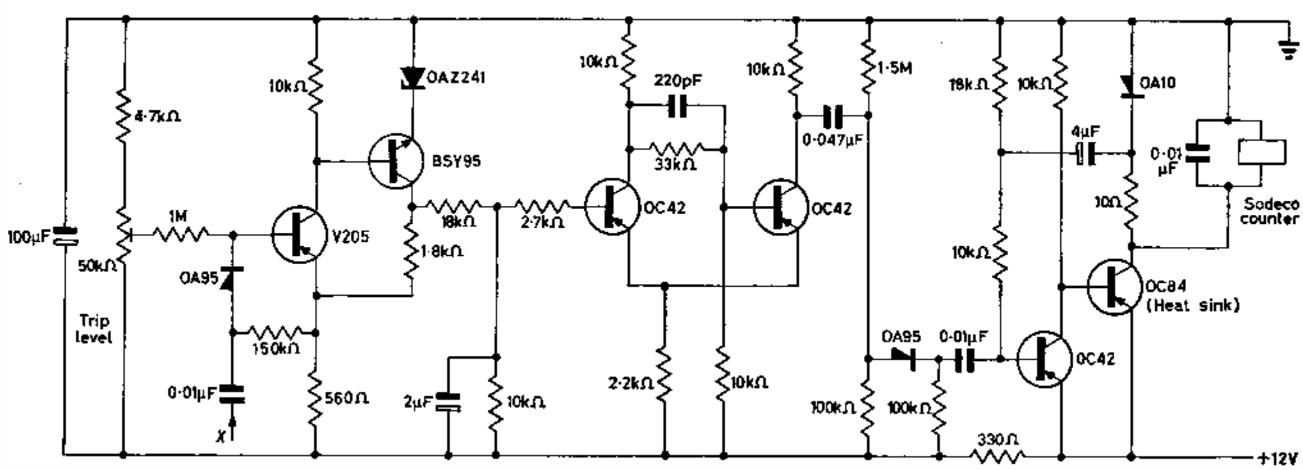
A 100c/s master multivibrator controls the basic repetition rate of the device; it triggers a silicon controlled rectifier, feeding a miniature ignition coil, to produce the 8kV pulse necessary to trigger the xenon flash tube light source. This tube produces 1μsec flashes at an energy of 0·04 joule. The power consumption of the flash circuit is minimized by charging the 0·05μF storage capacitor through a 200H ringing choke instead of the usual resistor with its associated power loss.

The 100c/s master multivibrator also triggers a 25μsec monostable circuit which pulses the fifth dynode of the photomultiplier to its normal operating potential. This very greatly increases the photomultiplier sensitivity for the duration of the 25μsec pulse.

After division by two, by a bistable circuit, the master multivibrator pulses trigger a further 25μsec monostable circuit which in turn grid-pulses the modulator valve. The valve produces 25μsec duration pulses of u.h.f. energy of about 100W peak at a repetition rate of 50 per second. The actual light modulation is effected by the a.d.p. crystal situated at a position of high electric field within the modulator cavity resonator.

The 1μsec returned light pulse detected by the photomultiplier, while it is gated 'on', sets the 30kc/s tuned circuit in its anode ringing, thus producing a train of damped oscillations and effectively stretching the pulse. The high dynamic impedance of the tuned circuit is matched to the detector by a complementary emitter-follower impedance convertor. The time-constant of the detector is such as to produce a saw-tooth waveform which is the envelope of the damped oscillations. This

Fig. 6. The counter circuit



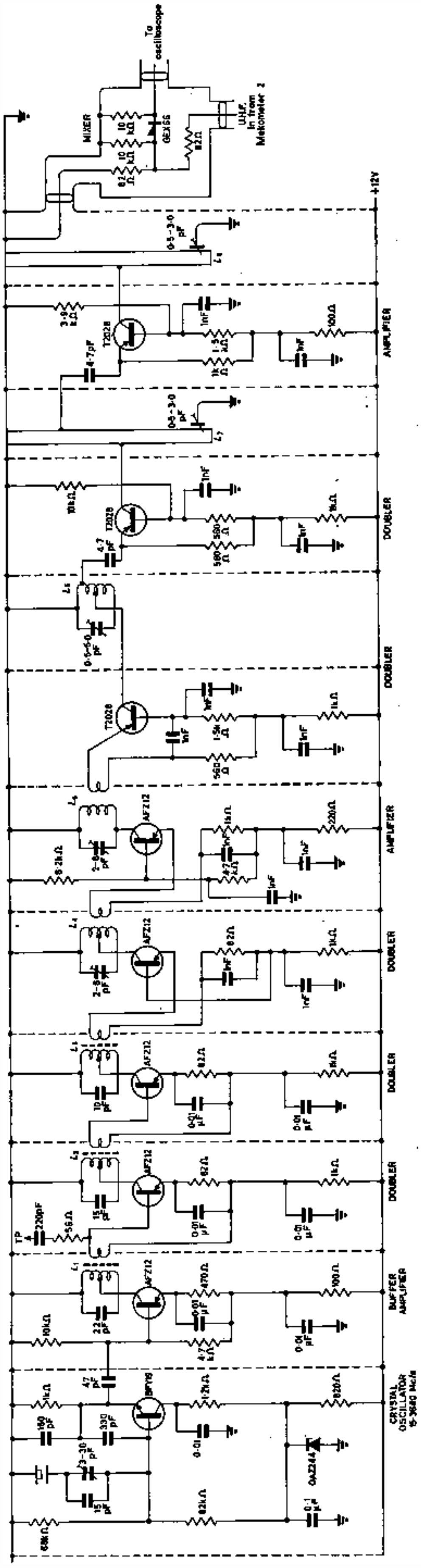


Fig. 7. The crystal-controlled frequency standard

is amplified by a two stage negative feedback amplifier and fed to a transformer, with a centre-tapped secondary winding, which produces two anti-phase outputs. These outputs are each rectified. A synchronous switch, activated by the bistable divider at 50c/s, selects each rectified output in turn and connects it to the null-indicating meter. Since the light signal is modulated on alternate flashes only, the synchronous system serves to eliminate unwanted signals detected by the photomultiplier.

The light intensity received by the photomultiplier can vary over an enormous range depending on the distance being measured and the prevailing meteorological conditions. To cope with this, gain control of the detector circuit has been achieved by varying the e.h.t. to the photomultiplier. Provision has also been made so that the gating pulse to the photomultiplier maintains a constant ratio to the e.h.t. voltage.

The modulation wavelength of the a.d.p. modulator is monitored by the two standard cavity resonators. These cavities are separated by a 5 per cent wavelength change. Changing the wavelength of the modulator between cavities *A* and *B* is achieved by a sliding capacitive ring within the modulator cavity, which can be moved by an electric motor and lead screw. The motor is automatically stopped at the cavity resonant positions by the circuit shown in Fig. 5.

The standard cavities, which have an unloaded *Q* factor of about 2 000, are very loosely coupled to the modulator. Some of the u.h.f. input is mixed with the phase-shifted signal that has been transmitted through the cavity so that, on resonance, a phase-balanced null is achieved.

The output from the respective cavity is detected and the resulting pulse stretched and amplified by a negative feedback amplifier. A second emitter-follower coupled detector rectifies the consequent signal and the resultant direct current is displayed on the resonance indicating meter.

A 0·2 per cent wavelength shift, which is also required by the measuring process, is achieved, without a third standard cavity, by means of an accurate, mechanical stop-controlled capacitive plunger in the modulation cavity wall.

The measuring process has two variants[3].

The first, and most satisfactory, involves three phase settings. One each at the resonance of the two standard cavities *A* and *B* and the third after introducing an exact 0·2 per cent wavelength shift (position *C*). A preliminary rough measurement, to the nearest 500ft, is achieved by shifting the modulator wavelength from one phase-null to the next and reading distance from a calibrated dial. This ensures no ambiguity up to 5 000ft.

The second method involves the phase settings at *A* and *B* standard cavity resonances as before. However, as the modulator is tuned between standard cavities *A* and *B* the revolutions of the phase pattern are counted. Each count represents 20ft, thus ambiguity is again resolved.

Fig. 6 shows this *n*-counter circuit. The signal applied to the phase-null meter is d.c. amplified and fed to a Schmitt trigger which in its turn operates a 20msec monostable circuit when the amplifier output reaches a predetermined d.c. level. The 20msec pulse switches an electromagnetic counter one digit, which represents 20ft.

The Mekometer II consists of two units, the first contains the accumulator and power convertors and weighs 15 lb, and the second the instrument proper plus its electronic systems and weighs 23 lb. The semiconductor circuits are contained on seven plug-in printed circuit boards. A flexible cable is used to connect the power unit to the instrument.

Several months experience with the Mekometer II suggest that the electronic systems employed are stable and trouble free over a reasonable temperature range. This range could be greatly increased by using silicon semiconductors throughout.

The pin based modulator valve is being run outside manufacturers' limits. It is hoped in the near future to re-design the modulator cavity to incorporate a ceramic disk-seal triode which will give a greater modulation index without being overated.

Initial Calibration

Since the Mekometer II system is independent of the velocity of light, a straightforward and most satisfactory method of adjusting the standard cavities is to set up the instrument on a known distance and trim the cavities so as to produce the correct answer.

A second and more involved method requires a knowledge of the vacuum velocity of light and the prevailing meteorological conditions.

The u.h.f. oscillations produced during the 25μsec pulse from the a.d.p. modulator are heterodyned with a c.w. signal produced by frequency multiplication from a fundamental quartz crystal oscillator at about 15Mc/s. The beat signal is viewed on an oscilloscope. Detection is made easy by the fact that the pulsed and c.w. oscillator phase-lock, a phenomena which is not readily explained.

This latter calibration method is shown in the circuit diagram of Fig. 7 and warrants a more detailed description.

The oscillator, which is of the Clapp-Gouriet type, uses a fundamental AT cut quartz crystal. The supply to the oscillator is Zener diode stabilized. No crystal oven has been employed since the frequency standard is normally used in a temperature controlled laboratory. After a common emitter buffer amplifier three stages of class-C doublers follow. The third doubler and all subsequent stages are in common base configuration. After the third doubler the signal is amplified back to a working level and two further doublers follow. The last doubler and the output amplifier are of trough-line construction.

The frequency standard has a temperature coefficient of 1 part in 4 $\times 10^6$/°C and changes 1 part in 2 $\times$ 10^6/V variation in supply voltage.

Both methods described for the initial calibration of the Mekometer II are in good agreement.

Future Developments

It is hoped to improve greatly the sensitivity of setting on an interference minimum and on the standard cavity resonances by introducing a small frequency modulation of the modulating cavity. This can be accomplished either mechanically, by means of a vibrating plunger, or electronically by means of a variable reactance placed inside the modulating cavity. By these means the modulating frequency will be changed in a cyclic manner between alternate light pulses. Thus both the interference minima and the standard cavity resonance will appear as meter 'zeros' and should give a greatly enhanced precision of setting. This should approach 0·1mm for distances up to one or two hundred metres, and 0·3mm thereafter.

A further modification which should give enhanced long-term stability, is to replace the invar centre conductors of the standard cavity resonators with ones of copper plated silica. These could extend the whole length of the cavity, to give mechanical rigidity, but only be plated part way up the inner conductor.

Acknowledgement

The Mekometer described has been developed as part of the research programme of the National Physical Laboratory and this article is published by permission of the Director.

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2. Bradsell, R. H. The NPL Mekometer. *Trans. Soc. Instrum. Tech.* *16*, 30 (1964).
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The following British patent applications are involved: 919368; 44480/63, 4564/64, 7473/64, 7687/64, 7858/64 and the cognated version of these.

A Microminiature Digital Computer

A general purpose microminiature digital computer suitable for use in civil and military aircraft or in similar difficult environmental conditions such as field use in civil engineering, military applications or geophysical prospecting has been developed by G.E.C. Computers and Automation Ltd (formerly International Systems Control Ltd).

This computer, known as the G719, is based on the earlier computer DEXAN (*D*igital *EX*perimental *A*irborne *N*avigation) used originally as a navigational aid.

It consists of a wide range of data processing equipment modules including principally the G719 Processor, the G729 Volatile Store Unit, the G739 Transfluxor Store and the G749 range of Input/Output Equipment.

A feature of this system is an external interface 'Highway' which permits the various units to be connected individually through common digital interfaces, thus providing considerable flexibility in assembling systems of different configurations.

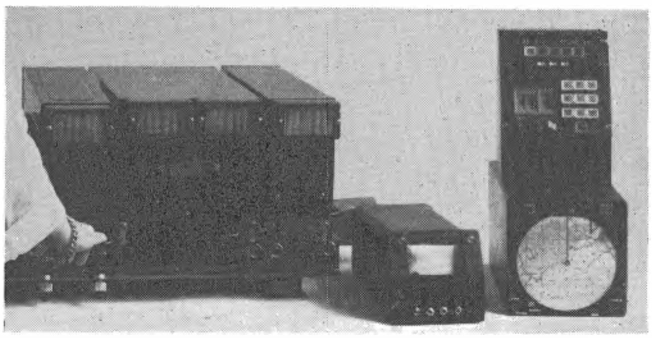
The G719 Processor, occupying less than 1/5ft^3 and weighing only 10 lb, is a fast parallel digital processor with a 20-bit word length (optionally 21 or 24 bit words) and an addition time of 8μsec including store access time.

The G729 Volatile Store Unit has a capacity of 8192 words, each word containing 20 bits, the store being based on the coincident current method of selection, using lithium ferrite cores for improved temperature performance. The cycle time is 2·5μsec.

The G739 Transfluxor Store is a random access 4096-word 20-bit store which also includes a 256-word volatile store and has a non-destructive read-out together with standard logical and electrical connexions to the store 'Highway'.

The range of G749 Input/Output Equipment consists of: Analogue to Digital/Digital to Analogue Convertors; Discrete Input Unit; Discrete Output Unit; Miniature Printer Unit; Teleprinter Unit; Tape Reader Unit; Tape Punch Unit.

A general purpose aircraft computer system comprising a processor, store and input/output equipment, occupying less than 1ft^3 and weighing 35 lb. To the right of the computer are a miniature Teleprinter, a time control panel and a moving map display.



An Improved RC Bridge Circuit with High Selectivity

By Sergio Bernstein*

An RC bridge circuit is presented in which the inclusion of an active unity-gain stage increases the effective circuit Q to a value 3/2 of that obtained with a standard Wien bridge. Such an improvement is especially beneficial in oscillator and selective amplifier circuits operating at audio or sub-audio frequencies.

(Voir page 344 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 351)

OF the group of basic frequency selective bridge circuits, the Wien bridge is the most often used since in practical applications it embodies the best compromise between ease of tuning over a wide frequency range and sharpness of phase-versus-frequency characteristic[1].

The twin-T and bridged-T networks also have their place in various applications, but a fourth network, sometimes referred to as the 'zero phase shift network'[2] has not had wide application in actual equipment.

This basic network is shown in Fig. 1(a), and there exists a similar network of identical characteristics with the elements R and C interchanged, as in Fig. 1(b).

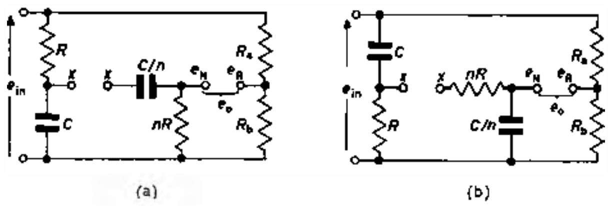


Fig. 1. Zero phase shift networks

The ratios of the magnitudes of the impedances in the reactive arms of all these bridge circuits can be adjusted to optimize their characteristics at null for any particular application.

It will be shown that the highest selectivity obtainable with a six element RC bridge circuit is equivalent to the selectivity of a bridged-T network with a $Q_o = 1$, whereas the usual Wien bridge configuration has an equivalent $Q_o = 2/3$.

This selectivity can be obtained with the circuits of Fig. 1 by letting the 'tapering' parameter n approach infinity. In practice, this can be accomplished by inserting a unity gain amplifier between the points x-x, whereas the inclusion of such buffing amplifiers is impractical in the ordinary Wien bridge circuit.

General Characteristics

All of the bridge networks shown in Table 1 are characterized by a transfer function, T, of the form†:

$$T = E_o/E_i = k\,\frac{1}{1 - j\,(b\rho/(\rho^2 - 1))} = k\,\frac{1}{1 - j\,(b/u)} \quad \ldots (1)$$

and by a phase characteristic, θ, of the form[3]:

$$\theta = \tan^{-1}\frac{b\rho}{\rho^2 - 1} = \tan^{-1}(b/u) \quad \ldots\ldots\ldots (2)$$

k is the transmission factor at null, and b is a phase factor. Let:

$$\rho = \omega/\omega_o \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

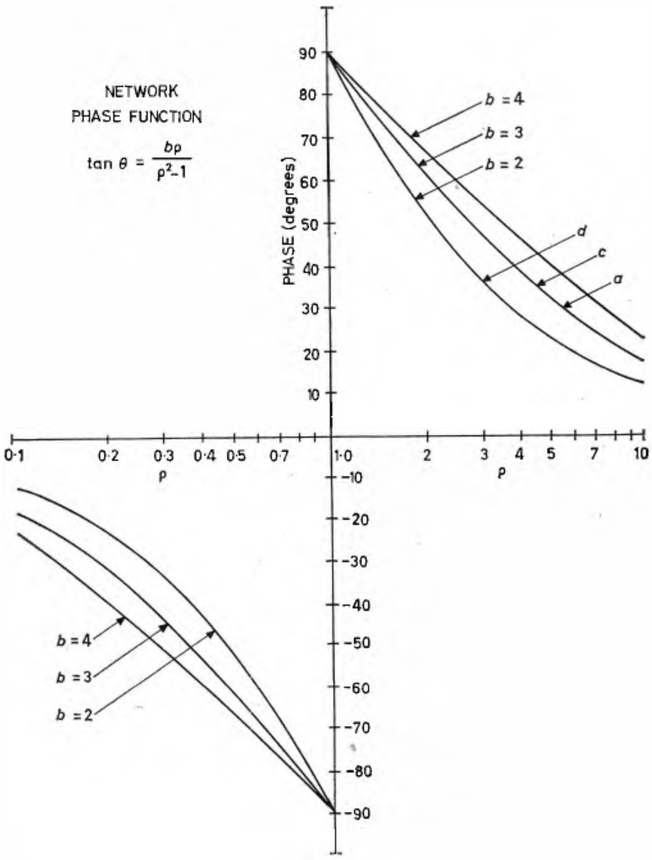


Fig. 2. Network phase characteristics

and:

$$u = \omega/\omega_o - \omega_o/\omega = \frac{\rho^2 - 1}{\rho} \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

Parameters k, b, and ω_o are given in Table 1, for various configurations.

The T networks have a phase characteristic which varies from 0 to $-\pi/2$ as ρ increases from 0 to 1 (ω increases from 0 to ω_o) and which varies from $+\pi/2$ to 0 as ρ increases from 1 to infinity (ω increases from ω_o to infinity). A discontinuity exists at $\rho = 1$, the phase changing from $-\pi/2$, through $+\pi$, to $+\pi/2$ degrees.

The bridge networks exhibit an additional π radians phase shift due to the negative sign associated with the transmission factor k.

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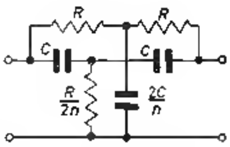
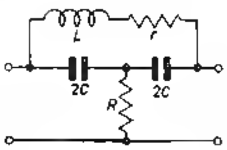
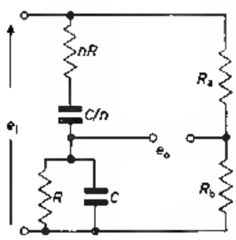
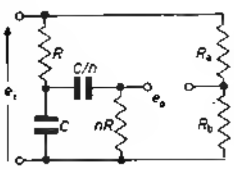
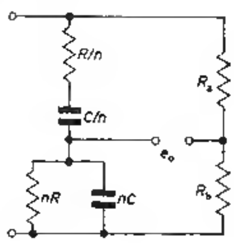
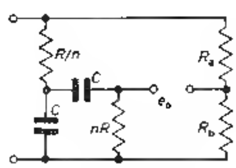
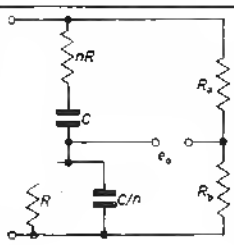
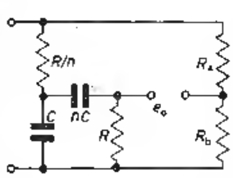
* *Perkin-Elmer Corporation, U.S.A.*

† *The transfer functions for the zero phase shift bridge are derived in a manner similar to that given in Volume 18 Op. Cit., p. 384 ff., for the Wien bridge, the bridged-T and twin-T networks. Another bridged-T network[4] using a resistance instead of an inductance in the circuit of Table 1, network (b), does not produce a true null and is therefore excluded from this discussion.*

TABLE 1

Network	Circuit	Condition for Null	k	b	$\lvert S_\theta \rvert$	S_T
Twin-T (a)		$\omega_0 = \frac{\sqrt{n}}{RC}$	$+1$	$2\left(\frac{n+1}{\sqrt{n}}\right)$ Min $b = 4$ for $n = 1$	$\frac{\sqrt{n}}{(n+1)}$ Max $S_\theta = 1/2$ for $n = 1$	$\frac{\sqrt{n}}{2(n+1)}$ Max $S_T = 1/4$ for $n = 1$
Bridged-T (b)		$\omega_0^2 = \frac{1}{LC}$ $R = \frac{Q_0 \omega_0 L}{4}$	$+1$	$\frac{2}{Q_0}$	$Q_0 = \frac{\omega_0 L}{r}$	$\frac{Q_0}{2}$
Wien Bridge (c)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b = 2n$	$\frac{-1}{1+2n}$ Max $k = -1$ for $n = 1$	$\frac{1+2n}{n}$ Min $b = 2$ for $n = \infty$ (3 for $n = 1$)	$\frac{2n}{1+2n}$ Max $S_\theta = 1$ for $n = \infty$ (2/3 for $n = 1$)	$\frac{n}{(1+2n)^2}$ Max $S_T = 1/8$ for $n = 1/2$ (1/9 for $n = 1$)
Zero Phase Bridge (d)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b = \frac{1+n}{n}$	$\frac{-n}{1+2n}$ Max $k = -(1/2)$ for $n = \infty$	$\frac{1+2n}{n}$ Min $b = 2$ for $n = \infty$ (3 for $n = 1$)	$\frac{2n}{1+2n}$ Max $S_\theta = 1$ for $n = \infty$ (2/3 for $n = 1$)	$\frac{n^2}{(1+2n)^2}$ Max $S_T = 1/4$ for $n = \infty$ (1/9 for $n = 1$)
Wien Bridge (e)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b = \frac{n^4+1}{n^2}$	$\frac{-n^2}{n^4+n^2+1}$ Max $k = -(1/3)$ for $n = 1$	$\frac{n^4+n^2+1}{n^2}$ Min $b = 3$ for $n = 1$	$\frac{2n^2}{n^4 n^2+1}$ Max $S_\theta = 2/3$ for $n = 1$	$\frac{1}{(1+n^2+(1/n^2))^2}$ Max $S_T = 1/9$ for $n = 1$
Zero Phase Bridge (f)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b = 2/n^2$	$\frac{-n^2}{n^2+2}$ Max $k = -1$ for $n = 1$	$\frac{n^2+2}{n}$ Min $b = 2\sqrt{2}$ for $n = \sqrt{2}$	$\frac{2n}{n^2+2}$ Max $S_\theta = 1/\sqrt{2}$ for $n = \sqrt{2}$	$\frac{n^3}{(n^2+2)^2}$ Max $S_T = \frac{3\sqrt{6}}{32}$ ($\simeq$ 0·23) for $n = \sqrt{6}$ ($\simeq$ 2·46)
Wien Bridge (g)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b =$ $(1/n)+n$	$\frac{-n}{1+n+n^2}$ Max $k = -(1/3)$ for $n = 1$	$\frac{1+n+n^2}{n}$ Min $b = 3$ for $n = 1$	$\frac{2n}{1+n+n^2}$ Max $S_\theta = 2/3$ for $n = 1$	$\left(\frac{n}{1+n+n^2}\right)^2$ Max $S_T = 1/9$ for $n = 1$
Zero Phase Bridge (h)		$\omega_0 = \frac{1}{RC}$ $R_a/R_b = \frac{1+n}{n^2}$	$\frac{-n^2}{1+n+n^2}$ Max $k = -1$ for $n = \infty$	$\frac{1+n+n^2}{n}$ Min $b = 3$ for $n = 1$	$\frac{2n}{1+n+n^2}$ Max $S_\theta = 2/3$ for $n = 1$	$\frac{n^3}{(1+n+n^2)^2}$ Max $S_T = \frac{(1+\sqrt{13})^3}{8(5+\sqrt{13})^2}$ ($\simeq$ .165) for $n = \frac{1+\sqrt{13}}{2}$ ($\simeq$ 2·31)

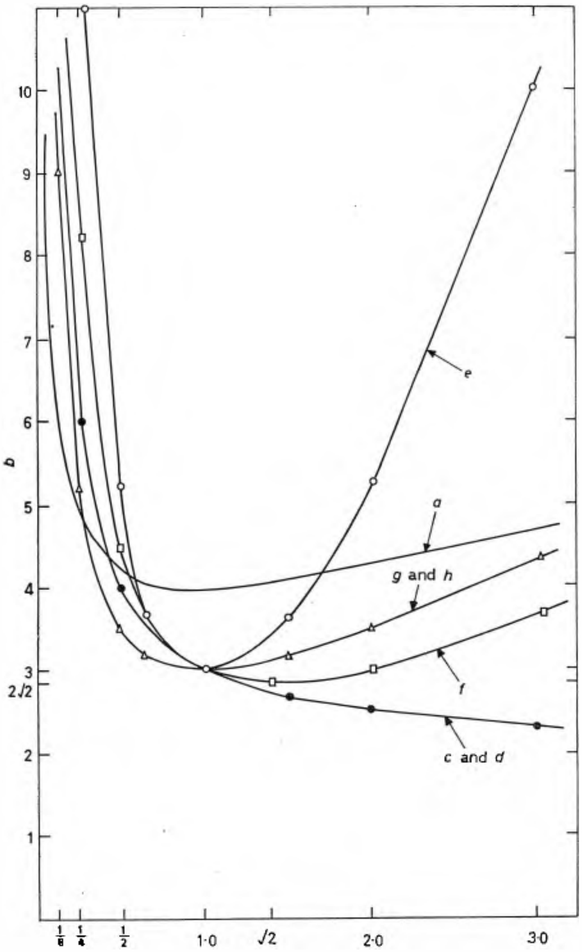


Fig. 3. Phase factor b of networks (Table 1)

Thus, for these networks, the phase characteristic varies from $-\pi$ to $-3\pi/2$ as ρ increases from 0 to 1, and from $\pi/2$ to π as ρ increases from 1 to infinity. Again a discontinuity occurs at $\rho = 1$, the phase changing from $-\pi/2$ through 0 to $\pi/2$ degrees.

The phase characteristics of the bridge networks have been plotted in Fig. 2 neglecting the extra-π radians phase shift associated with the coefficient k. Thus, the rate of change of phase with frequency of the T and bridge networks can be more easily compared.

Selectivity

A third and most important parameter, which describes the behaviour of these networks is their selectivity factor[5], S_θ, where:

$$S_\theta = \frac{d\theta}{d\omega/\omega_0} \quad \ldots\ldots (5)$$

It is the ratio of the differential rate of change of phase with respect to differential rate of change of frequency around the null frequency.

The largest possible selectivity factor is desirable. This will minimize the deviation from the true null frequency due to residual phase shifts in associated amplifiers, when these circuits are used in actual oscillator circuits[6].

Sensitivity

Another parameter, sometimes referred to in the literature[7] is the sensitivity factor* at null, S_T, where:

$$S_T = \left.\frac{d\,|T|}{d\omega}\right|_{\omega=\omega_0} = \left.\frac{d\,|T|}{d\rho}\right|_{\rho=1} = \left.\frac{d\,|T|}{du}\right|_{u=0} \quad \ldots (6)$$

It is the ratio of the differential rate of change of output voltage with respect to differential rate of change of frequency at the null frequency.

Maximizing this particular parameter is of importance when these networks are used to measure resistance or capacitance rather than frequency, since a bridge null detector is usually sensitive to changes in output voltage magnitude only, and not to phase changes.

The selectivity factor for all the bridge networks under consideration is found to be†:

$$S_\theta = (-2/b) \quad \text{or:} \quad |S_\theta| = 2/b \quad \ldots\ldots\ldots\ldots (7)$$

their sensitivity factor is given by‡

$$S_T = k/b \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (8)$$

thus:

$$S_T = (-k/2)\,.\,(S_\theta) \quad \ldots\ldots\ldots\ldots\ldots (9)$$

The tapering parameter n can be varied for each configuration to maximize either S_θ or S_T.

Either factor can be maximized for each particular application. In four configurations, a, d, e, and g of Table 1, the selectivity factor, and sensitivity factor, S_T, maxima occur simultaneously. Note that the transmission factor, k, is then also at its maximum value.

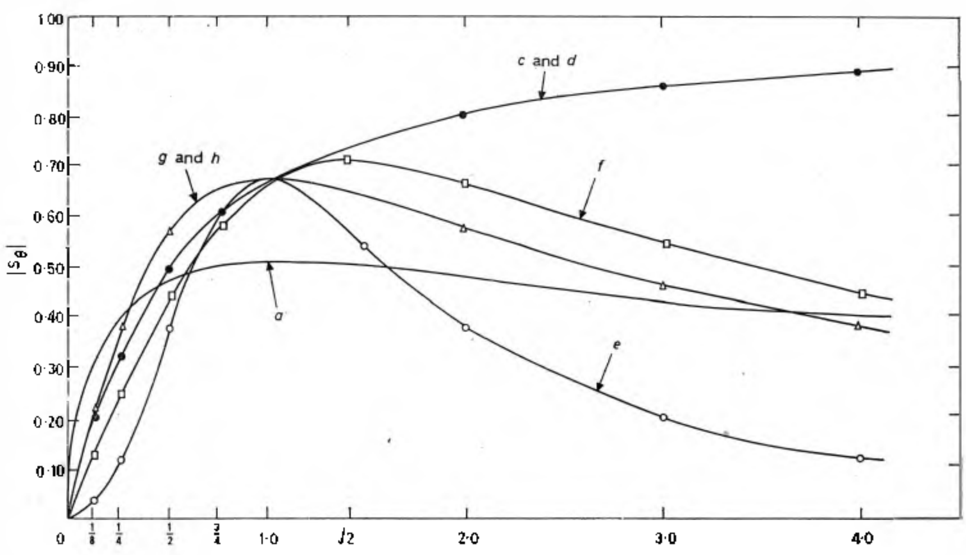
Furthermore, if n is made infinite by using a buffer amplifier as shown in Fig. 7(a), $b = 2$ and the maximum possible selectivity is realized, since each RC section gives precisely 45° phase shift (one leading, the other lagging) and is not loaded by the second section. Thus, at the null frequency, each section operates at the point where both the rate of change of phase and of amplitude, with change of frequency, are the greatest.

* *Sometimes this parameter is also referred to as 'selectivity' which often leads to confusion*[3].

† *Derived in Appendix (1).*

‡ *Derived in Appendix (2).*

Fig. 4. Selectivity characteristics of networks (Table 1)



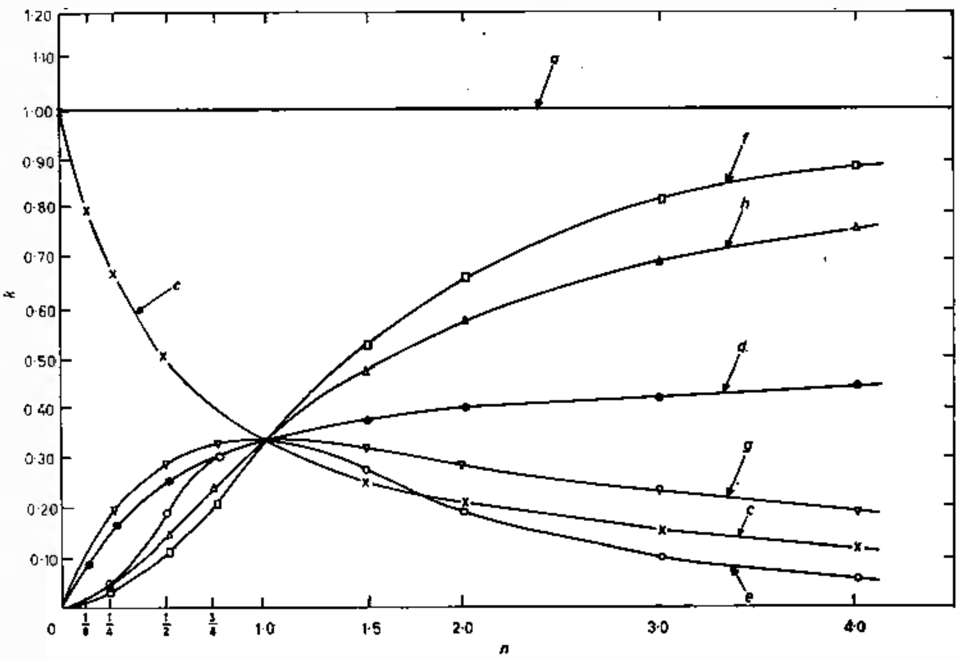


Fig. 5. Transmission characteristics of networks (Table 1)

The selectivity is then 3/2 that of an ordinary Wien bridge, and equivalent to that of a bridged-T circuit with $Q_o = 1$.

In practice, a buffer amplifier having a high input impedance, a low output impedance and unity voltage gain, is required. A cathode-follower or (compound) emitter-follower is well suited in this application.

Optimum Characteristics

It is interesting to compare the selectivity characteristics of the T and bridge networks, as a function of the parameter n, since selectivity is of prime importance in oscillator and selective amplifier applications.

It is readily apparent that maximum selectivity is obtained when the phase factor b is a minimum. The minimum value of b for the twin-T network is 4 for $n = 1$, indicating that it is less selective than the other bridge configurations.

The Wien bridge of Table 1(c) has a phase factor $b = 3$ for $n = 1$, while a minimum value for b is obtained with the zero phase bridge of (d). When $n = \infty$, $b = 2$ which is the smallest phase factor obtainable, unless a bridged-T configuration is used with $Q_o > 1$. See Fig. 2.

Curves of the various factors as a function of the tapering parameter n are plotted in Figs. 3 to 6.

The selectivity of the Wien bridge of Table 1(c) can also be increased by increasing n, but only with a concomitant decrease in transmission factor k. Thus as this circuit is made more selective, higher amplifier gains are required. For maximum selectivity, $k = 0$ and the circuit becomes impractical.

For the zero phase bridge of Table 1(d) however, an increase in n causes an increase in k. Thus, a network is obtained in which the selectivity increases while its attenuation decreases for increasing values of n!

For $n = \infty$, $k = 1/2$, and the transmission of this network is 3/2 that of the standard Wien bridge.

The transfer function of this new network is:

$$T = -1/2 \cdot \frac{1}{1 - j\,(2\rho/(\rho^2-1))}$$

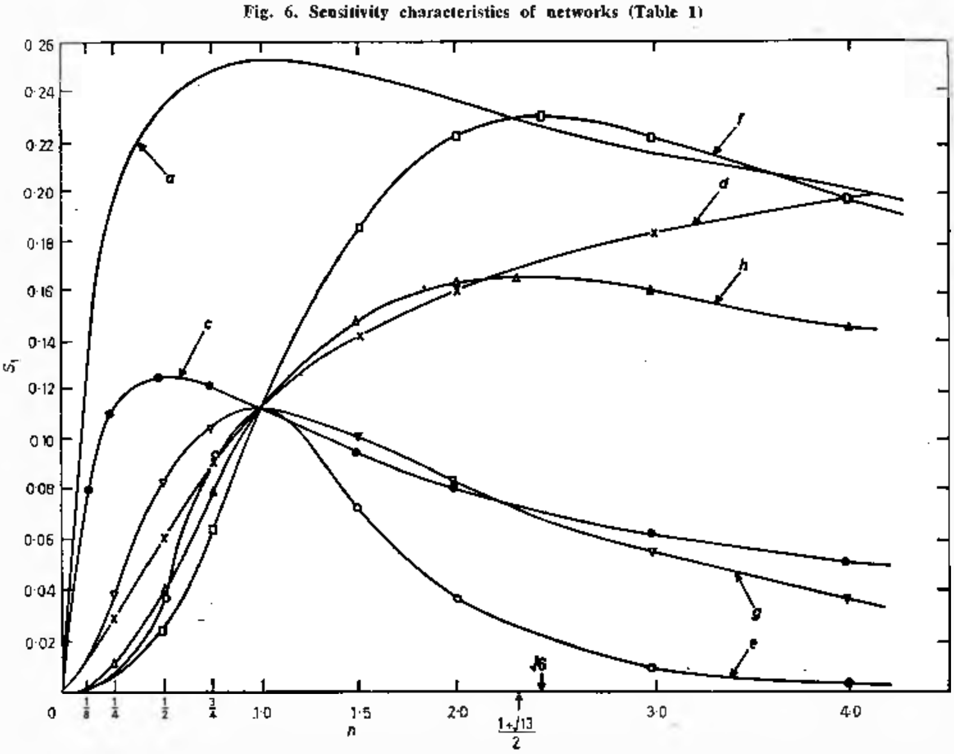
$$= -1/2 \cdot \frac{1}{1 - (j2/u)} \quad .. \ (10)$$

It is interesting to note that an output voltage having a phase angle of precisely 45° (with reference to the input or output voltages) is always available from this network, regardless of the value of n. This feature is sometimes very useful[9].

Consider now the sensitivity characteristic[7,8], S_T, plotted in Fig. 6.

Circuit (c) of Table 1 has a maximum sensitivity for $n = 1/2$. Thus, for this circuit, sensitivity and selectivity are incompatible since to maximize one the other is sacrificed.

Fig. 6. Sensitivity characteristics of networks (Table 1)



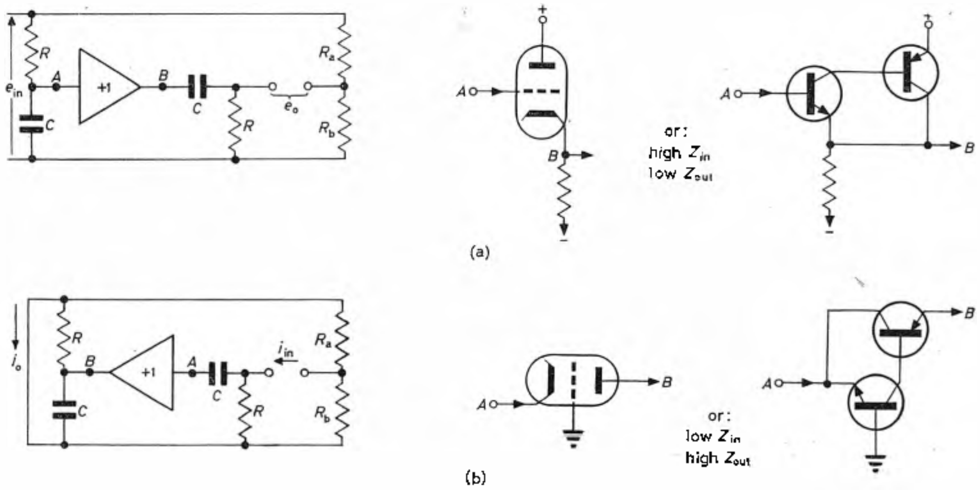


Fig. 7. (a) Optimized zero phase shift bridge network and suitable buffer amplifiers
(b) Current dual of optimized zero phase shift bridge network and suitable buffer amplifiers

Circuit (h) exhibits maximum sensitivity for $n = \frac{1 \pm \sqrt{13}}{2}$ $\simeq 2{\cdot}305$, without a great reduction in maximum selectivity. Circuit (f) has the greatest sensitivity of any configuration, excepting circuit (d) in which n is made infinite by the addition of a buffer amplifier. Its selectivity factor is less than that of circuit (d) for any $n \geqslant 1{\cdot}2$ so that again, some sacrifice in selectivity must accompany an optimized sensitivity characteristic.

Current Dual Network

The usefulness of the results derived previously are not limited to the voltage dependent networks presented, but are readily applicable to current dependent duals of such networks. The current dual is sometimes a more desirable configuration when used with transistors[10].

Since for passive networks[11] $g_{21} = -g_{12}$ the results also apply when each network is driven in the opposite direction by a current generator at the output terminals, with the input terminals shorted. See Fig. 7(b). The buffer amplifier required must now have a very low input impedance, a high output impedance, and unity current gain. A grounded grid amplifier or (compound) grounded base amplifier is well suited in this application.

Conclusions

A new, simple bridge network has been found, whose selectivity factor and transmission factor are 3/2 times, and its sensitivity factor 9/4 times that of the standard Wien bridge circuit.

These improved factors lead to the realization of the following benefits and improvements:

(1) For a given oscillator stability factor, and output waveform distortion, an amplifier having only 2/3 the open loop gain of the amplifier used for a standard Wien bridge circuit is sufficient.

(2) Since the attenuation of the new network is only 2/3 that of the Wien bridge network, a further 2/3 open loop gain reduction is permissible in the amplifier.

(3) Since a gain reduction of $(2/3)^2$ is permissible in the open loop gain of the associated amplifier, the oscillator frequency range could be extended by a factor of $(3/2)^2$ at high frequencies.

Only two negligible inconveniences appear to accompany the considerable benefits described above:

(a) The two capacitors no longer have a common terminal.

(b) For any given lamp used as a stabilizing element (R_b), the voltage output from the new bridge circuit, will be $\frac{2}{3}$ that of the Wien bridge circuit, since it is desirable to operate these lamps around a definite, fixed, optimum point.

Appendix

(1) Selectivity

The selectivity factor, S_F, is defined as:

$$S_\theta = \frac{d\theta}{d\omega/\omega_0} \quad \ldots\ldots\ldots\ldots (11)$$

It should be as large as possible for maximum selectivity and stability of the associated circuit.

For the bridged-T, twin-T, Wien bridge, and zero phase bridge networks, the transfer function is of the form:

$$T = k \,.\, \frac{1}{1 - j\,\dfrac{b\rho}{\rho^2 - 1}} \quad \ldots\ldots\ldots\ldots (12)$$

and the phase characteristic is of the form:

$$\theta = \tan^{-1} b\left(\frac{\rho}{\rho^2 - 1}\right) = \tan^{-1} b\left(\frac{\omega \,.\, \omega_0}{\omega^2 - \omega_0^2}\right) \equiv \tan^{-1} u \quad \ldots\ldots\ldots\ldots (13)$$

where $\rho = \omega/\omega_0$

Letting:

$$u = \frac{b \,.\, \omega \,.\, \omega_0}{\omega^2 - \omega_0^2} \quad \ldots\ldots\ldots\ldots (14)$$

Then:

$$d\theta/d\omega = \frac{1}{1 \,.\, u^2} \,.\, du/d\omega$$

and:

$$d\theta/d\omega = -\frac{(\omega^2 + \omega_0^2)\, b\omega_0}{[\omega^2 - \omega_0^2]^2 + [b\omega_0\omega]^2} \quad \ldots\ldots\ldots\ldots (15)$$

To find the selectivity factor S_F:

$$S_\theta = \omega_0 \,.\, \frac{d\theta}{d\omega}\bigg|_{\omega=\omega_0} = -\frac{[b\omega_0\omega_0 \,.\, \omega_0^2 + \omega_0^2]}{[b\omega_0\omega_0]^2}$$

Simplifying:

$$S_\theta = (-2/b) \text{ or: } |S_F| = (2/b) \quad \ldots\ldots\ldots (16)$$

which is the general expression for the selectivity factor of the bridge networks given in Table 1.

(2) Sensitivity

The sensitivity factor, S_T, is defined as:

$$S_T = \frac{d\,|T|}{d\omega}\bigg|_{\omega=0} = \frac{d\,|T|}{d\rho}\bigg|_{\rho=1} = \frac{d\,|T|}{du}\bigg|_{u=0}$$

For all the networks in Table 1, the transfer function, T, is of the form:

$$|T| = k\,\frac{1}{1 - j\,\dfrac{b\rho}{\rho^2 - 1}} = k\,\frac{u}{u - jb} \qquad (17)$$

and:

$$|T| = k\,\frac{u}{(u^2 + b^2)^{1/2}} \qquad (18)$$

Differentiating equation (18) with respect to u:

$$d\,|T| = k\left[\frac{1}{(u^2 + b^2)^{1/2}} - \frac{u^2}{(u^2 + b^2)^{3/2}}\right] \qquad (19)$$

Letting $u = 0$ in equation (19) to find the sensitivity factor at null:

$$\left.\frac{dT}{du}\right|_{u=0} = k/b$$

Therefore:

$$S_T = k/b \qquad (20)$$

This factor is a function of n, as given in Table 1, and can be optimized for each configuration by letting: $dS_T/dn = 0$, solving for n, and using this value of n in equation (20).

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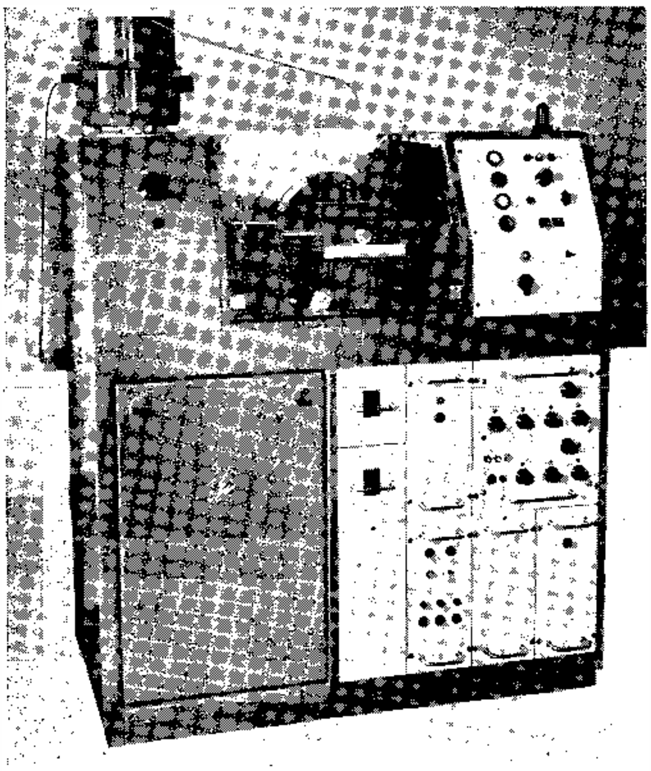
An Automatic Spiral Grinding Machine

A machine which grinds the spiral in metalized or carbon film resistors and simultaneously measures the resistance value has been developed by VEB Electromat of Dresden, East Germany and is available through the U.K. agents Techna (Great Britain) Ltd.

The resistor bodies are fed singly into the grinding machine from a hopper by the conventional vibratory methods and the first step in the automatic process is to measure the 'unground' or body resistance. If this is within ±15 per cent of the nominal value, the machine proceeds to grind a spiral on the resistor body while it is being rotated.

During this process the resistor body forms one arm of a Thomson bridge and the grinding continues until the resistance reaches the nominal value when the resistor is released and another unground resistor is fed into the machine. The whole process is automatic and the machine continues to grind resistors at the rate of three a minute.

The spiral grinding machine



The pitch of the spiral is adjusted so that nominal resistance value is reached by grinding to between 70 and 100 per cent of the body length. If the nominal value is reached before 70 per cent of the resistor body length is ground or the value is not reached when the full length is ground, the resistor body is rejected.

The occasional incorrect resistance is allowed to be rejected, but if more than ten resistors are rejected consecutively the machine shuts down and a warning signal is given. The machine will also stop if the hopper is empty.

The machine will grind resistor bodies of lengths of 6mm to 20mm and diameters of 1·8mm to 6mm.

With body resistance values ranging from 1Ω to 100kΩ spiral ground resistors can be produced between 100Ω and 10MΩ.

When the nominal values of body and final resistances are correctly set into the machine, the accuracy of the spiral ground resistor depends on the dimensions of the resistor body, the rotational speed of the resistor and the grinding pitch, but an accuracy of 0·3 per cent can be attained.

'Crossbar' Telephone Switching

The Post Office has announced that the Plessey '5005' Crossbar system will be used in some of its large new telephone exchanges.

Although small (up to 2 000 line) electronic exchanges are now in service the need for faster switching, greater flexibility and reliability still remains. however, for new large exchanges and in order to bridge the gap until the major joint developments in large electronic switching systems reach a practical stage, the Plessey '5005' Crossbar system will be used to meet some of the Post Office requirements.

Plessey '5005' Crossbar has been designed to interwork with all existing telephone switching systems and it can readily be adapted if necessary to extend economically and gradually replace Strowger.

This modern system, the only one in its class being made in Britain, was originally developed primarily for export, and financed from the Company's own resources, '5005' Crossbar has achieved many successes overseas, despite fierce international competition.

Apart from the obvious advantages of faster operation and low maintainence costs, '5005' Crossbar has a number of other important technical merits. For example, each Crossbar switch is built into a small self-contained unit, together with its own control circuit which operates independently of other units. This allows the 'unit construction' principle to be applied throughout the system, thus enabling each switching item to be manufactured, wired, shipped, installed and maintained as a separate entity.

The '5005' Crossbar system is manufactured by Automatic Telephone and Electric Co. Ltd, a member of the Plessey Telecommunication Group.

Trailing Edge Variable Slope Pulse Modulation Circuits

By O. E. Kruse* and J. K. Kincaid*

The electronic circuits for the production and detection of a train of modulated triangularly shaped pulses are described. The intelligence to be transmitted is carried in the slopes of the trailing edges of the train of pulses. In the demodulation process, the variable slope pulses are converted to variable amplitude pulses, which pulses are then amplified, stretched, and filtered. The final output is a good reproduction of the original intelligence.

(Voir page 344 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 351)

A PULSE modulation system employing a train of equally spaced, equal amplitude, triangular shaped pulses carrying intelligence in the slopes of the leading edges of the pulses has been described[1] as has the spectrum analysis of the modulated train of pulses[2]. The method has been further described and has also been called pulse-slope modulation[3,4] or p.s.m. In this article the circuits for a modulation system employing a train of equally spaced, equal amplitude, triangular shaped pulses carrying intelligence in the trailing edge of the train of pulses, and henceforth called trailing edge variable slope pulse modulation, will be described.

Trailing Edge Modulator

The basic arrangement of the trailing edge modulator is shown in Fig. 1, and its operation will be described briefly. The battery has very low internal resistance, and so the load does not affect its output voltage appreciably. When the switch closes, the capacitor charges rapidly, and after a very short time the switch opens and allows the capacitor to discharge through the current regulator which allows a current, i, to flow. This process is repeated periodically with a period of T_c.

The current through the current regulator is:

$$i = i_0 + G\,a(t) \quad \ldots\ldots\ldots\ldots\ldots\ (1)$$

where i_0, G, and $a(t)$ are current with no modulating signal, proportionality constant, and intelligence, respectively.

The equation for the voltage across the capacitor, and so the output, is:

$$V(t) = V_b - (1/C)\int i\,dt \quad \ldots\ldots\ldots\ldots\ (2)$$

The slope of the pulse is:

$$dV(t)/dt = -i/C \quad \ldots\ldots\ldots\ldots\ (3)$$

which becomes:

$$dV(t)/dt = (-1/C)\,[i_0 + G\,a(t)] \quad \ldots\ldots\ldots\ (4)$$

by substitution from equation (1).

In actual operation the sample time intervals of $a(t)$ are very short (of the order of microseconds); therefore, the sampling voltage is almost a constant and so provides a 'linear' slope for each trailing edge. Of course, the slopes of successive pulses are dependent upon the values of $a(t)$ at the times of samplings.

Fig. 2(a) shows the trailing edge modulator circuit. A rectangular pulse generator is used as the capacitor charging unit. The rectangular pulse generator allows variation of the frequency, amplitude, and duration of the rectangular pulse train. An npn transistor which has a relatively constant collector current for wide ranges in collector voltages with constant base current is placed in parallel with the capacitor. To understand the circuit operation, let the output voltage of the rectangular pulse generator and the output of the modulator be zero when observation begins. The modulator output voltage continues to be zero until the rectangular pulse applies a voltage across the capacitor, which charges rapidly. The

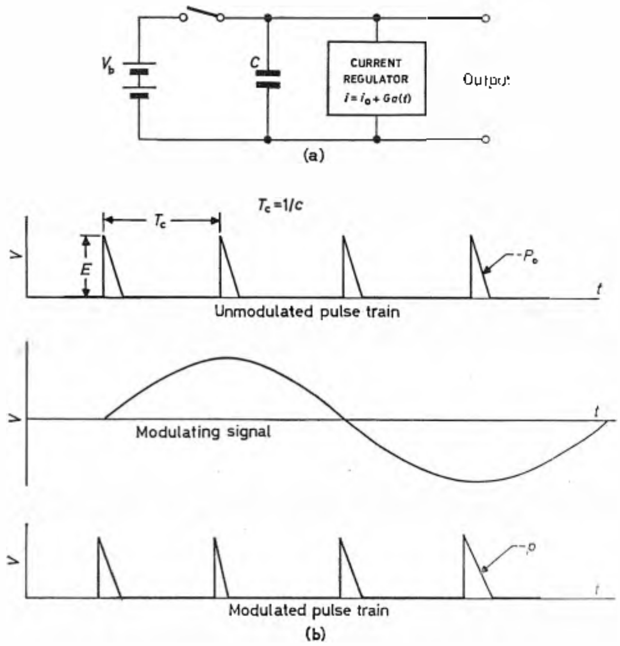


Fig. 1. Arrangement and operating sequence of trailing edge slope modulator

rectangular pulse is short, and the output of the rectangular pulse generator again becomes zero. The discharge path of the capacitor is through the transistor, for the diode will not allow current to pass into the cathode-follower circuit of the rectangular pulse generator; therefore, the capacitor discharges at a constant rate depending on the base current of the transistor. Another sample is taken when another positive pulse comes from the rectangular pulse generator. The base current of the transistor is controlled by the modulating signal, and so the slope of the trailing edge varies with the signal applied to the base. A diagram of the operation of this unit is shown in Fig. 2(b). The degree of modulation, m, can be changed easily by varying the input signal to the transistor base; the unmodulated slope, $-P_0$, can be varied by changing the base current of the transistor; E can be changed by varying the amplitude of the rectangular pulse generator; furthermore, c can be varied by varying the pulse repetition rate of the rectangular pulse generator.

Trailing Edge Demodulator

To demodulate the trailing edge variable slope pulse modulated wave, the wave train is first inverted by a

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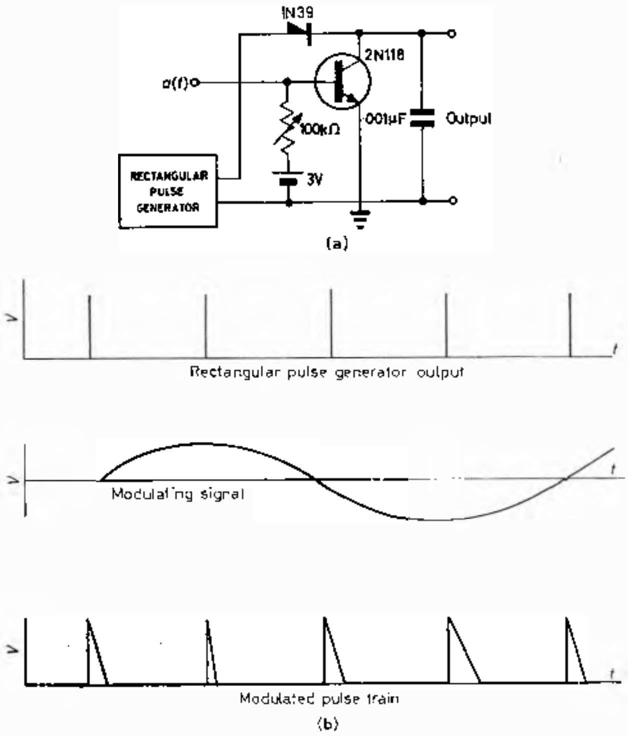

Fig. 2. Trailing edge slope modulator system
(a) Schematic diagram of trailing edge slope modulator
(b) Operating sequence

suitable and simple phase invertor and then fed across the input to the circuit shown in Fig. 3. The circuit of Fig. 3 is the one used quite succesfully in demodulating leading edge variable slope pulse modulation. After inversion of the trailing edge modulated pulse train, the demodulator works equally as well as a demodulator of the trailing edge variable slope pulse modulation wave.

The key to the operation of the demodulator is the simple series C_dR_d circuit. If the product C_dR_d is chosen much smaller than the rise (or decay) times of the pulses, the voltage across R_d reaches a peak value that is proportional to the slope of the pulse. This fact can be established easily, and has been employed in some related work[5]. Thus, since the slopes are proportional to the modulating signal voltage at the times of the samplings, it follows that the peaks of the voltage pulses across R_d are now directly proportional to the modulating signal voltage at the sampling times. The variable slope pulse train has then been converted by the C_dR_d circuit into a variable amplitude pulse train.

The remainder of the circuit in Fig. 3 simply amplifies, stretches, and filters the variable amplitude pulses so that the original modulating signal with sufficient energy will be available at the output. A brief explanation of the circuit will be given. Valves V_1 and V_2 and accompanying circuits serve as pulse amplifiers. The output of V_2 drives the two cathode-followers, V_3 and V_4. The output of cathode-follower V_4 is applied to the anode of the diode, V_5, which conducts when a pulse arrives and so charges the 500pF capacitor in its cathode circuit. The capacitor remains charged because the triode V_6 is normally biased below cut-off. However, the output of cathode-follower V_3 is used to trigger the multivibrator, consisting of V_7 and V_8, which, by means of its flip-flop action, causes a sharp positive pulse to be applied to the grid of V_6 at some later time and prior to the appearance at the input of another of the modulated pulses. The sharp positive pulse on the grid of V_6 drives the valve into strong conduction, with a consequent rapid discharge of the 500pF capacitor. The capacitor is then ready to be charged by another pulse, probably of different amplitude. The procedure of charge and later discharge is repeated for successive pulses.

The train of wide, variable amplitude pulses is applied to the grid of the cathode-follower, V_9. The output of the cathode-follower is applied through a 10kΩ resistor to a low-pass audio filter having a cut-off frequency of 3 500c/s. The output of the filter is applied to the grid of an audio frequency amplifier, V_{10}, amplified and made available at the secondary of a suitable audio transformer in the anode circuit of the amplifier, thus reproducing the original intelligence, $a(t)$.

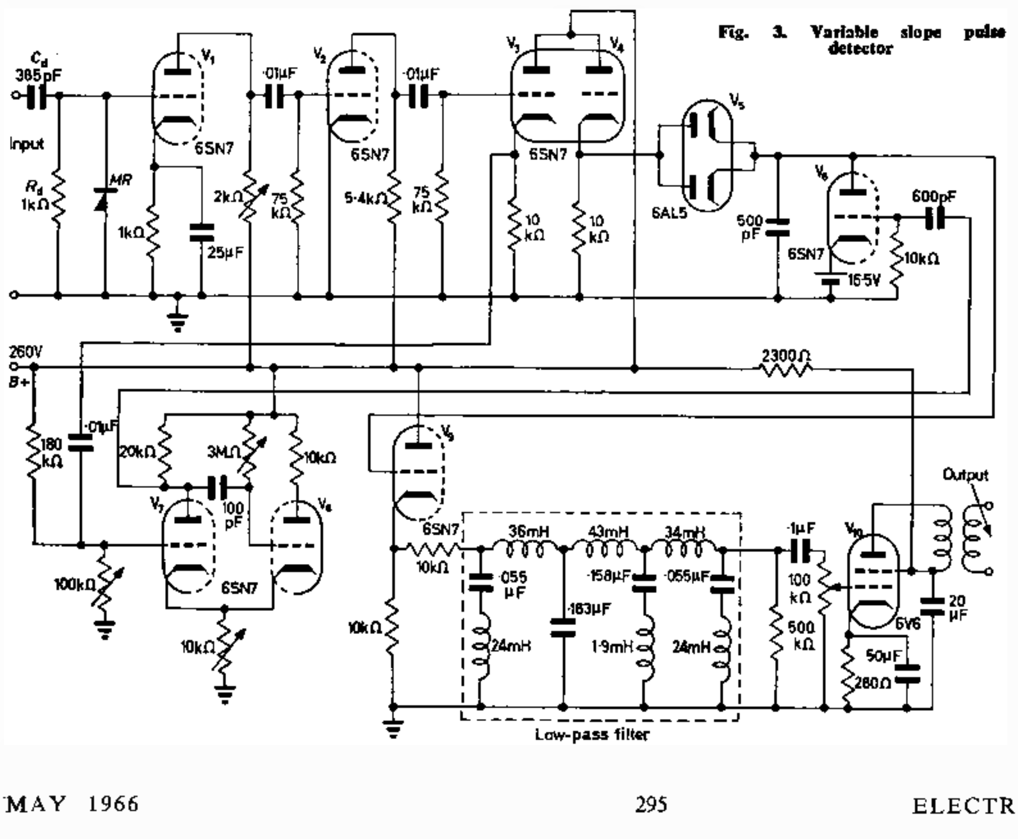

Fig. 3. Variable slope pulse detector

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A Versatile Digital Frequency Synthesizer for use in Mobile Radio Communication Sets

By A. F. Evers*

This article describes a tunable precision frequency source which uses a phase locked variable frequency oscillator. The closed loop frequency control system, which includes digital fixed and variable ratio frequency dividers, compares a variable sub-multiple of the output frequency from the v.f.o. with a fixed sub-multiple of the output from a high stability frequency standard. This 'digital frequency synthesizer' is shown to be suitable for a wide range of applications in mobile radio sets where small size, low power consumption and high reliability are important. An extremely pure output can be obtained, at the expense of extra complexity and slightly increased power consumption, if adequate precautions are taken to isolate the slave oscillator both from the frequency control system and from the effects of environmental disturbances. Some of the more important design problems are discussed and the characteristics of several laboratory and production prototype models of the digital synthesizer are described.

(Voir page 345 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 352)

INCREASING use of the h.f. and v.h.f. bands of the radio spectrum has stimulated considerable activity in the field of frequency synthesis during the past five years. Frequency synthesizers are already in common use for static transmitting and receiving installations, but the bulk and complexity of most synthesizers has so far inhibited their widespread use in mobile radio sets. Recent improvements in semiconductor devices, and the advent of microelectronic circuits, have removed most of these restrictions. It is now possible to make a synthesizer that is sufficiently small and reliable for it to be used in lightweight man-carried equipments as well as in aircraft and vehicle installations.

For these mobile applications the operational demands are such that a synthesizer of quite unusual performance is required to meet them. The synthesizer must generate an extremely pure output signal and, in addition to being small and reliable, it must also consume very little power. Nevertheless its cost must not be disproportionate in relation to the cost of the complete radio set.

To investigate the best means of meeting these requirements a research programme was started in 1961. The object was to devise a method of frequency synthesis which was compatible with microelectronic circuit techniques. It was clear even then that a system which used mainly digital circuits would have considerable advantages over existing synthesizers of the type which included many filters, harmonic generators and heterodyne mixers. The basis of a purely digital synthesizer was proposed by J. D. Newman[1] and a working model was first demonstrated in 1961. Since that time the digital frequency synthesizer has been developed in several versions covering frequency bands up to 400Mc/s. More recently miniature synthesizers, which occupy less than 10in^3 and use only 1W of primary supply power, have been made, using semiconductor and thin film integrated circuit elements.

The development programme has also emphasized the considerable versatility of this type of synthesizer. Because all the circuits are aperiodic, most circuit modules can be made common to a wide range of synthesizer applications covering frequency bands between 1 and 400Mc/s. The digital synthesizer is not limited to specific construction techniques—microelectronic or conventional component assemblies can be used to satisfy particular compromises on cost and performance.

The aim of this article is to describe briefly the basic principles of the digital frequency synthesizer, to review its advantages and to discuss how it can be integrated into radio equipments.

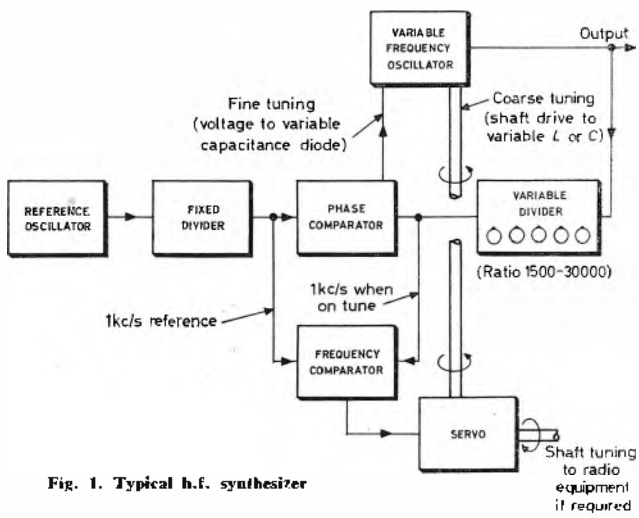


Fig. 1. Typical h.f. synthesizer

Digital Frequency Synthesizers

Method of Synthesis

The principles of digital frequency synthesis can be explained by considering the simplified block schematic diagram of a typical h.f. synthesizer (Fig. 1). The output is obtained from a phase-locked slave oscillator which is tunable over the required frequency range. The oscillator can be coarsely tuned by components which are adjusted or selected mechanically, for example by a servo motor or by switches. The frequency can also be finely adjusted electronically by means of a voltage sensitive reactance such as a variable capacitance diode (v.c.d.).

In the frequency control unit a variable sub-multiple of the oscillator frequency (f_o) is compared with a fixed sub-multiple of a standard frequency (f_{ref}). The error signal from a frequency comparator is first used to control coarse tuning of the slave oscillator; then when the frequency falls within the capture range of the phase lock loop, the error signal from the phase comparator causes the oscillator to phase lock to the required frequency. When this occurs the inputs to the comparators are equal in frequency and the output frequency is given by:

$$f_o = (V/F) \,.\, f_{ref}$$

* *The Plessey Co. Ltd.*

where V and F are the division ratios of the variable and fixed ratio frequency dividers. The comparison frequency equals f_{ref}/F, which in this case is 1kc/s. The division ratio V can be changed in unit steps between 1 500 and 30 000, corresponding to an output frequency range of 1·5Mc/s to 30Mc/s covered in increments of 1kc/s. The most important component of the frequency control system is clearly the variable ratio frequency divider (v.r.f.d.) whose ratio setting determines the controlled frequency. In a typical synthesizer it occupies over half the volume and is by far the major component in terms of circuit complexity.

The v.r.f.d. is basically a digital counter. It consists of several decade or sub-decade counter stages, each of which corresponds to one digit of the output frequency. The decades in turn are built up from a range of standard logic circuits which includes a J-K flip-flop and NOR/NAND gates. The decades use parallel logic for high speed operation[2] and can operate either in a straight divide-by-ten mode or in a 'reset' mode, starting from any one of ten preselected states. Feedback connexions between cascaded decade stages are used to obtain the required overall count. The division ratio is preset by applying a binary-coded pattern of d.c. inputs to 'reset' gates associated with each counter element. The required pattern is produced by the frequency setting switches.

An important feature of the variable counter is its ability to divide by a number which differs by a constant amount from the number which corresponds to the frequency indicated on the frequency setting switches. This facility, which will be referred to as 'side-step', permits the synthesizer to generate the local oscillator frequency for a receiver while the setting switches indicate the aerial frequency. Sidestep can be permanently wired in, or it can be switched in on command from a separate switch such as a transmit-receive switch.

In a practical synthesizer the simple arrangement described above is usually modified by including a further fixed ratio divider between the slave oscillator and the variable divider. The need for this additional divider arises because it is difficult to design variable counting circuits which will operate at the maximum output frequency. This is especially true for counters constructed from integrated circuit elements. At the present time reliable operation of solid circuits, with power consumption which is low enough for man-carried portable radio sets, is limited to frequencies below about 10Mc/s. For unrestricted power consumption the upper frequency limit is about 25Mc/s.

Introducing a fixed ratio divider into the control loop has some disadvantages, however. The frequency at which phase comparison takes place is reduced and so the bandwidth of the low-pass filter must also be reduced. (The function of this filter is to remove unwanted comparison frequency components from the phase error signal). In general both the value of the comparison frequency and the filter characteristic determine the response of the control system Thus a reduced comparison frequency results in less control over short term fluctuations of the oscillator frequency, and an increase in the time taken to achieve phase lock. In practice it has proved desirable to operate the phase comparator at or above 250c/s. For an h.f. synthesizer this allows the inclusion of a divide-by-four counter before the variable divider, thus reducing the maximum input frequency to the variable divider to about 8Mc/s. The characteristics of this type of phase lock loop are somewhat unusual. Because phase comparison takes place at a fixed frequency the capture range of the system is always a constant fraction of the output frequency. A capture range of 5 per cent is quite easily achieved and consequently coarse tuning of the oscillator need not be very accurate.

The control loop includes the variable divider which has a transfer function whose magnitude varies inversely with the controlled frequency. The transfer function of the slave oscillator must compensate for this variation since it is desirable for the loop gain to be independent of frequency. In the ideal case the fractional change of frequency per volt should be independent of frequency. The implications of this requirement will be discussed later.

Advantages of Digital Synthesizers

The digital frequency synthesizer was selected in the first instance because the volume available in most applications could only be achieved by using microelectronic techniques. This type of synthesizer, being com-

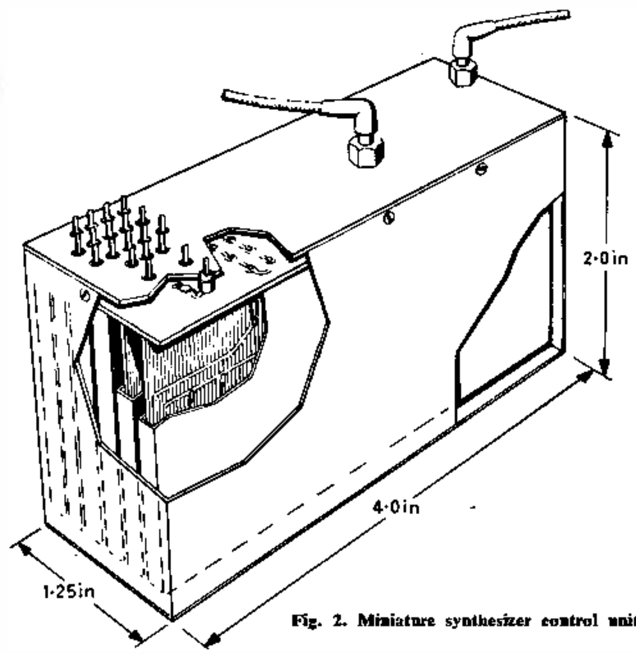


Fig. 2. Miniature synthesizer control unit

pletely aperiodic, requires no filters or other bulky components. Most of its counter stages can be constructed using standard digital semiconductor integrated circuits (s.i.c.'s). Thin film circuits or densely packed discrete components can generally be used for linear circuits such as buffer amplifiers and for high speed fixed ratio counters. A digital synthesizer constructed with s.i.c.'s and some thin film and discrete component circuit modules occupies between 10 and 20in³, depending on its application. For high stability h.f., s.s.b. equipments the volume quoted refers to the synthesizer 'control unit' and does not include either the slave oscillator or the frequency standard (the reasons for dissociating the slave oscillator from its control system are explained later). For a v.h.f., f.m. set covering 35 to 57Mc/s the complete frequency generator, including a medium stability frequency standard, occupies about 8in³. The use of microelectronic circuits is expected to give a considerable improvement in reliability and ultimately to reduce the cost of manufacture.

In addition to the advantages which accrue directly from the use of solid circuits the digital synthesizer has other advantages. Synthesizers operating in the h.f., v.h.f. and u.h.f. bands are largely constructed from common parts. A common v.r.f.d. usually operates at a maximum frequency of about 10Mc/s and the frequency range of the controlled oscillator is extended to higher frequencies by including extra fixed ratio counters in the control loop.

The two comparators and the reference counter are also common to all applications. Clearly the synthesizer is extremely flexible in this respect, although some of this flexibility is exchanged for some restriction on the minimum frequency increment that can be achieved for each frequency band. This is because of the limitation on comparison frequency mentioned previously—in general the minimum channel spacing must increase as the maximum controlled frequency increases. However, the limitation on minimum frequency increment already exists for other reasons (for example the stability of reference oscillators). At present, typical values of increment are 1kc/s between 1Mc/s and 30Mc/s, 12·5kc/s between 30Mc/s and 100Mc/s and 25kc/s between 100Mc/s and 400Mc/s.

The h.f. synthesizer already described includes a divide-by-four counter in the control loop. By including a further divide-by-four counter in the loop a nearly identical system can control a v.h.f. oscillator covering 30 to 100Mc/s in 12·5kc/s increments. For 12·5kc/s steps one stage of the v.r.f.d. is modified for a radix of 8 instead of 10 and the comparators operate at 781·25c/s (i.e. 12·5kc/s divided by 16) instead of 250c/s.

Another important characteristic of the digital frequency synthesizer is its very low power consumption. For applications in man-carried transmitter-receivers this is most important; military users often require that a 12V, 2Ah secondary battery should be capable of supplying a complete radio set for 12 hours on a single charge. This means that the synthesizer must consume less than 1W. In the digital synthesizer most circuits operate in a switching mode at frequencies below 1Mc/s. Hence low power saturating logic circuits operating with very low voltage supplies can be used.

Finally a very pure output is obtained from the digital synthesizer. A pure output is characteristic of any synthesizer which uses a phase locked oscillator as its output frequency generator, providing adequate precautions are taken to isolate the oscillator from its control system. Some of these precautions are described later.

TABLE 1

Module description	Class of radio equipment: H.F./S.S.B. man-pack	V.H.F./F.M. vehicle	V.H.F./U.H.F./A.M. airborne	maximum input frequency
Slave oscillator	3·750 to 31·749Mc/s	39·6875 to 109·6625Mc/s	150·000 to 349·975Mc/s	
Buffer amplifier	BUFFER	BUFFER	BUFFER	32, 110 or 350Mc/s
High speed fixed ratio frequency dividers			÷ 2	350Mc/s
		÷ 4	÷ 4	180Mc/s
	÷ 4	÷ 4	÷ 4	45Mc/s
Variable ratio frequency divider	HF DECADE	HF OCTADE	HF QUARTADE	11Mc/s
	DECADE	DECADE	DECADE	4Mc/s
	DECADE	DECADE	DECADE	
	DECADE		DECADE	
	QUARTADE	DUODECADE	QUARTADE	
Phase and frequency comparators	PHASE	PHASE	PHASE	250c/s or 781·25c/s
	FREQUENCY	FREQUENCY	FREQUENCY	
Fixed ratio frequency dividers	÷ 200	÷ 200	÷ 200	160kc/s
	÷ 105	÷ 31	÷ 32	5Mc/s
	BUFFER	BUFFER	BUFFER	5Mc/s
Frequency standard	5·25Mc/s	4·84375Mc/s	5Mc/s	

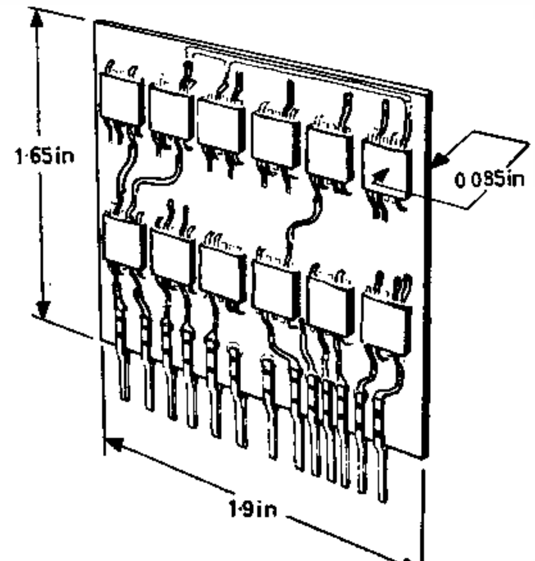


Fig. 3. Medium speed decade counter

Design Problems

Assembly and Packaging Techniques

Fig. 2 shows an assembled miniature synthesizer control unit and Fig. 3, a standard, medium speed decade counter —the basic building block of a variable divider.

Twelve s.i.c.'s, encapsulated in 8-lead flat-packs, are mounted on a printed wiring board measuring 1·8in × 1·4in. The Kovar ribbon leads of the s.i.c.'s are welded to the board which is clad with a nickel foil specially designed for welding. Joints are made by a parallel gap resistance welder using feedback from the welding probe to control the energy dissipated in the weld.

The synthesizer contains twelve modules, five of which are similar to the decade counter in that they are constructed entirely with solid circuits. The remaining modules use either a mixture of solid circuits and discrete components, or, as in the case of the phase comparator, they are constructed entirely with miniature discrete components. The use of discrete components is necessary at the present time for circuits which require close tolerance passive components, large value capacitors, or transistors with especially good low noise or high frequency characteristics.

Each of the modules is separately encapsulated in a protective resin and housed in a screening can which also serves as a mould for the encapsulation. The twelve modules are interconnected by a printed circuit 'mother board' and the complete assembly is housed in an outer case which provides additional screening and mechanical support.

Use of Common Modules

Table 1 shows how a few basic modules may be interconnected to form synthesizers for the following communications equipments:

(a) An h.f., s.s.b. man-pack or vehicle mounted transmitter receiver covering 2 to 30Mc/s in 1kc/s steps, with an i.f. sidestep of +1·75Mc/s.

(b) A vehicle mounted v.h.f., f.m. set covering 30 to 100Mc/s in 12·5kc/s steps, with an i.f. sidestep of +9·6875Mc/s.

(c) An airborne v.h.f./u.h.f., a.m. set covering 100 to 156Mc/s and 225 to 400Mc/s in 25kc/s steps, with an i.f. sidestep of ±50Mc/s.

In order that the minimum number of modules may be used most modules cover a frequency range which embraces the needs of all three applications. For example the fixed ratio divide-by-four counter, which operates between 3·25 and 31·75Mc/s for the h.f. version, also acts as the second divide-by-four counter in v.h.f. versions. Here the frequency range required is only 7·5 to 27·5Mc/s. Medium speed decade counters in the variable divider are common to all versions, although counters with non-standard radices are often needed for the first and last stages. An aperiodic 'sample and store' phase comparator is used—only three capacitors are changed to adapt this circuit from one version to another. The frequency comparator is also common and is constructed from standard s.i.c.'s. The fixed divider following the reference oscillator uses a chain of s.i.c. counter-register elements. Ratios other than 2^n are obtained by feedback.

Reduction of Power Consumption

A complete circuit diagram of an h.f. synthesizer would show that it contained some 800 transistors—most of them in about 70 solid circuits. Clearly, careful choice of the s.i.c.'s is the first step to achieving low power consumption. During the development programme several ranges of commercially available digital s.i.c.'s were examined. Most of the early circuits were rejected because of their poor power-speed product. More recent circuits using epitaxial and buried N+ techniques have the improved performance required. At present a series of circuits, constructed from combinations of a modified d.c.t.l. NOR gate[3], is used for all modules operating below 5Mc/s. This circuit uses saturating transistors and operates satisfactorily from a 3V supply at a consumption of 2·5mW per NOR function. The worst case propagation delay is 40ns. A counter-register element requires six of these NOR gates which are diffused into a single semiconductor chip and encapsulated in an eight lead package. Five register elements are used in a standard decade counter which consumes 120mW. The present performance of d.c.t.l. circuits is determined by transistor stored charge and by the parasitic capacitances of the integrated circuit. When techniques for making smaller solid circuit transistors and reducing parasitic capacitance by means of dielectric isolation are perfected, the performance should improve still further.

Apart from the s.i.c.'s the main users of power are the high frequency counters and buffer amplifiers used in the early stages of the synthesizer. The operating currents of these circuits inevitably increase with frequency because of the fundamental limitations imposed by stray capacitances and charge storage in transistors. Maximum current change and minimum voltage change are essential for high frequency switching or amplification. Power requirements are therefore minimized by using transistors with low values of stored charge and collector-base capacitance in simple circuits which operate with low supply voltages. Some of these circuits are described in the next section.

Methods of Synthesizing Very High Frequencies

There are two possible ways of modifying the basic synthesizer so that it may generate frequencies up to 400Mc/s. Either a heterodyne frequency convertor or an additional fixed ratio counter may be introduced into the control loop. The frequency convertor, although superficially attractive, has been avoided for two reasons:

(a) It is difficult to make microcircuit versions of the tuned frequency multiplier and mixer that would be required.

(b) The capture range, measured as a percentage of the output frequency, is reduced and is no longer independent of frequency.

Considerable effort has therefore been directed to designing counting circuits for reliable operation at frequencies up to 400Mc/s. The problems are mainly the familiar ones of designing for a military environment. The circuits must be tolerant of a wide range of operating conditions, including temperatures between −40°C and +100°C, and 10 per cent supply voltage variations. The need for circuits which consume minimum power greatly restricts the choice available. A high speed version of the Eccles-Jordan binary counter is satisfactory up to 100Mc/s. For higher frequency operation tunnel diode binary counters are used. These circuits are driven by short pulses, less than 1ns wide, which are generated by charge storage or 'snap-off' diodes. Table 2 summarizes the performance of the Eccles-Jordan and the tunnel diode frequency dividers.

Control of Spurious Emissions and Unwanted Modulation

Unwanted outputs from the synthesizer can be separated into four classes:

(a) Those corresponding to modulation of the wanted output and appearing as sidebands close to and associated with the carrier. Because the output is obtained from a phase locked oscillator, frequency modulation will be produced by spurious l.f. voltages superimposed on the controlling phase error

TABLE 2
Performance of high speed fixed ratio frequency dividers

CIRCUIT DESCRIPTION	MINIMUM FREQUENCY (Mc/s)			MAXIMUM FREQUENCY (Mc/s)			PERMISSIBLE SUPPLY VOLTAGE RANGE (V)	PERMISSIBLE R.F. DRIVE VOLTAGE RANGE (V r.m.s.)	POWER CONSUMPTION (mW)
	−40°C	+25°C	+100°C	−40°C	+25°C	+100°C			
Modified V.H.F. Eccles—Jordan ÷ 4 Counter	10	10	10	115	140	120	3±10%	1±20%	110 mW
V.H.F. Tunnel Diode ÷ 4 Counter	10	10	10	140	150	140	3± 5%	1±20%	100 mW
U.H.F. Tunnel Diode ÷ 2 Counter	10	10	10	350	360	350	20±10%	2+20%	50 mW

signal. Similar voltages superimposed on the oscillator h.t. supply can produce amplitude modulation.

(b) Discrete frequency outputs, not directly associated with the carrier. Equipment specifications generally require the amplitude of these components to decrease as the frequency separation between the unwanted output and the carrier increases.

(c) Wideband noise.

(d) Harmonics.

The main source of frequency modulation is the comparator frequency ripple voltage which is present on the phase control line. This is usually attenuated by an *RC* low-pass filter. The use of a sample and store phase comparator[5] simplifies the filter design because its ripple content is inherently low and because it does not introduce an extra lag into the control loop. A recent version of this phase comparator provides a control range of 3V (corresponding to 360° phase variation) with a ripple content of about 1mV peak-to-peak. An oscillator with a 5 per cent capture range has a deviation sensitivity of about 1c/s μV^{-1} at 30Mc/s. For h.f., s.s.b. speech transmissions a peak deviation of 10c/s is permissible at comparison frequency rate and hence the low-pass filter is required to attenuate comparator ripple voltages by more than 40dB.

Small random fluctuations of the phase error voltage can also result from semiconductor $1/f$ noise originating in the phase comparator, and from changes in the trigger potentials of counting circuits. The frequency deviation resulting from these causes can be kept below 1 part in 10^6 by using filtered h.t. supplies and low noise transistors for the appropriate circuits. A further improvement in the deviation due to $1/f$ noise is obtained by using a bistable phase comparator instead of the sample and store comparator. In the bistable comparator transistors operate in a switching mode and, being either saturated or cut-off, generator no $1/f$ noise.

The final source of unwanted frequency modulation is microphony and its effects will be discussed later.

Discrete frequency spurious outputs result from unwanted r.f. voltages appearing at the interconnexions between the slave oscillator and the frequency control unit. The main sources of interference are the fixed and variable counters which generate transients at sub-multiples of their input frequencies, and harmonics of these sub-multiples. Fortunately the energy content of these transients is small because of the low operating power. The h.t. interconnexions can therefore be isolated with fairly simple conventional filters. The main signal interconnexion however requires a high grade buffer amplifier to isolate the slave oscillator from the control unit. This amplifier has a very high attenuation in the reverse path between input and output, and low terminal impedances to avoid 'pick-up'. A practical amplifier with a reverse attenuation of 80dB at frequencies up to 100Mc/s has been constructed using two common-base stages and an output stage with collector-base feedback. This circuit arrangement reduces spurious voltages at the input terminal of the amplifier to less than 1μV r.m.s.

To complete the isolation of the slave oscillator from the control system, the oscillator, buffer amplifier and supply filters are each enclosed in screened boxes. This prevents pick-up of interfering signals due to radiation and magnetic or electrostatic induction.

Prediction of Reliability

The usefulness of the digital synthesizer in a wide range of military mobile communications equipments has already been discussed. Because of the increasing complexity of this type of radio set it has become necessary to take special precautions to ensure that the design and development process leads to a reliable end-product. The user now requires some kind of quantitative assurance that his equipment will continue to function for a useful period without failure. Ideally the average failure rate of the equipment should be predicted from statistical measurements made on standard production units. The result is usually expressed in terms of a 'mean time between failures' or m.t.b.f. For complex equipments with a low failure rate the process of providing statistical assurance of a predicted m.t.b.f. can be lengthy and expensive. An alternative which is acceptable in practice, is to predict the m.t.b.f. by using the best available information on the failure rates of individual components and processes. In this way the choice of components and processes can be controlled during the design stage.

In the case of components (including s.i.c.'s) to be used in the synthesizer, information on failure rates will be obtained from the manufacturers own reliability assurance programme. Processes, such as assembly, (e.g. welded joints), etching of printed circuit boards and encapsulation will also be evaluated at the design stage. In addition to the assessment of component failure rates both modules and complete units will be subjected to the environmental tests described in section L3 of the military specification DEF133. These tests are designed to reveal failure mechanisms which might occur in service and which can then be corrected during development. In the military applications of the digital frequency synthesizer one is concerned only with the frequency control unit and hence in this case the 'synthesizer' does not include either the slave oscillator or the frequency standard.

There are several design features of the frequency control unit which suggest that it is inherently reliable.

(a) The use of 70 solid circuits to replace about 1 000 discrete components considerably reduces the component and interconnexion complexity.

(b) About 90 per cent of the circuits are digital. Most modules consist of assemblies of a limited range of s.i.c. logic elements and each element is used many times in a complete synthesizer. System design is based on the worst case performance of the s.i.c.'s to provide adequate safety margins for circuit operation.

(c) The unit is required to operate over the temperature range −37°C to +100°C in its most stringent application. Components rated for temperatures between −55°C and +125°C are normally used. Since power consumption is low there are no serious heat transfer problems and because the power dissipation is evenly distributed there are no 'hot spots'.

(d) There are no electromechanical components in the control unit. In many versions of a complete synthesizer the only mechanical components required are the frequency setting switches and the oscillator tuner.

(e) The control unit is entirely constructed from components which are individually of low mass and which are firmly fixed by their lead-outs and encapsulation. Thus the probability of failure due to mechanical stress is low.

(f) Finally the control unit will, in many applications, be used inside sealed equipments. This reinforces the protection already provided by the use of sealed modules, against the ingress of moisture.

The reliability target for the frequency control unit is an m.t.b.f. of at least 1500 hours under operational conditions.

Predictions based on current estimates of component failure rates suggest that the actual m.t.b.f. is likely to be considerably longer than 1500 hours.

An approximate assessment of the m.t.b.f. for the control unit is given in Table 3. The failure rate figures used will be seen to be very conservative, particularly for the conditions of use described above.

The total failure rate for a complete assembly is predicted to be 23·2 per cent/1 000h. This converts to an m.t.b.f. of 100 000/23·2h or 4 000h approximately.

It is thought that the predicted value of m.t.b.f. could be in error by factors of 2 down or 4 up, putting the figure expected as a field result in the range 2 000 to 16 000h.

There are eleven modules in this particular control unit. The average m.t.b.f. per module is therefore predicted to be about 45 000 hours, equivalent to an average of about five years continuous operation without failure. Demonstration of a strong possibility of meeting this m.t.b.f. figure will therefore fully justify a philosophy of maintenance by replacement and discard of faulty modules.

Integration Into Radio Equipments

Co-operation with Radio Set Designers

Several military radio equipments are currently being developed as a joint exercise between two engineering groups. The first group has responsibility for the transmitter-receiver proper: the second group provides a digital synthesizer for each equipment. When an equipment is sub-divided in this way, it is usually convenient for the radio set designer to provide the slave and reference oscillators, as well as the frequency setting switches. The last two items are naturally the concern of the set designer, but the reason for dissociating the slave oscillator from the synthesizer is perhaps not so obvious. It is simply that most communications transmitter-receivers are tuned by ganged *LC* assemblies. The oscillator is intimately associated in the radio set with other tunable r.f. circuits which are tuned by a common shaft. The r.f. tuner must therefore be designed both mechanically and electrically as a single assembly.

In this situation the 'synthesizer' is no longer a frequency generator but rather a frequency control unit. It accepts inputs from the slave and reference oscillators, together with information from the frequency setting switches, and generates the error signals which control the tuning of the slave oscillator.

When the synthesizer is split in this way it is important to specify accurately the properties of both the oscillator and its control system and to define the conditions existing at the interfaces between them. The capture range of the phase lock loop, the loop gain and phase stability margins, and the purity of the oscillator output all depend on careful control of the following factors:

(a) The transfer function of the slave oscillator and the phase loop filter (the two are usually physically close together to avoid 'pick-up'). This will have the general form:

$$df/dv = fK_1 \frac{(1+j\omega T_1)}{(1+j\omega T_2)\,.\,(1+j\omega T_3)\,\ldots\,(1+j\omega T_n)}$$

where:

$T_1 \ldots T_n$ are the phase loop filter time-constants (for systems with a capture range of less than 5 per cent, $T_1 = 0$)

K_1 is the oscillator sensitivity, i.e. fractional frequency change per control volt

ω is the angular frequency of phase error voltage variations.

The amplitude constant, K_1 and the time-constants, $T_1 \ldots T_n$ must each have permissible maximum and minimum values specified.

(b) The transfer function of the control unit:

$$dv/df = j\,(K_2/f)\,.\,(\omega_\phi/\omega)$$

where:

K_2 is the phase comparator sensitivity (volts/radian)

ω_ϕ is the comparison frequency.

Again the tolerance on K_2 must be specified.

(c) The amplitudes of both wanted and unwanted signals at the r.f. interface between the oscillator and its control unit.

(d) The amplitude of unwanted l.f. signals present at the phase control interface.

(e) The amplitude of spurious voltages superimposed on supply voltage interconnexions.

Tuning Systems

In co-operation with several equipment manufacturers the synthesizer has been adapted to be compatible with four different methods of equipment tuning. These are:

(a) Manual or servo tuning using ganged variable capacitors. This method is most commonly used in the h.f. band in conjunction with a v.c.d. for electronic frequency control.

(b) Coarse tuning in discrete steps using switched capacitors selected by the frequency setting switches.

(c) Pure electronic tuning using v.c.d.'s to cover the whole of the required frequency range. V.H.F. bands up to half an octave wide can be covered in this way, the oscillator being tuned sequentially by a combined frequency and phase error signal[4]. Once phase lock is established no frequency error exists and the loop gain must therefore be sufficient for the phase error to hold the oscillator at the required frequency. The equivalent 'hold-in' range is 1·4:1.

(d) Manual or servo tuning using ganged variable inductors. These are usually used in v.h.f. and u.h.f. equipments and may take the form either of a permeability tuner or a shaft-driven variable inductor

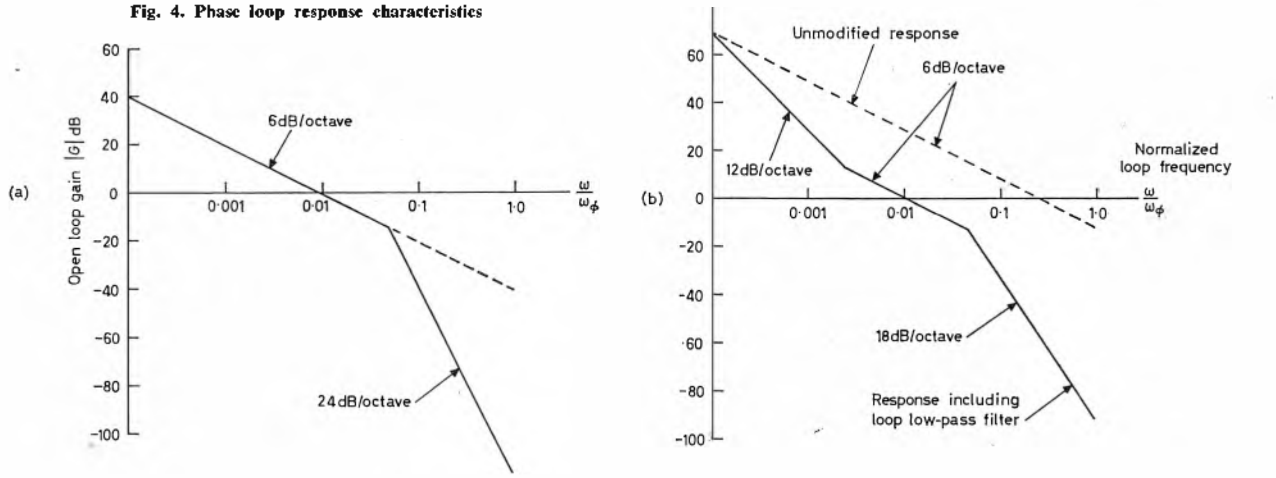
The characteristic of the tuning system most affected by the needs of the frequency control system is the magnitude of the oscillator transfer function (i.e., the fractional frequency change per control volt). This function should change as little as possible with frequency in order to avoid variations in control loop gain, which could result in unstable operation. Since the v.c.d. is commonly the

TABLE 3
Component Failure Rates

ITEM	NUMBER IN CONTROL UNIT	ASSUMED FAILURE RATE (per cent/1000h)	TOTAL FAILURE RATE OF ITEMS (per cent/(1000h)
Silicon Integrated Circuits	71	0·2	14·2
Transistors	27	0·1	2·7
Diodes	24	0.1	2·4
Resistors	63	0·025	0·16
Capacitors	38	0·025	0·1
Inductors	1	0·05	0·05
Joints	1000	0·003	3·00
Printed Circuit Boards	12	0·05	0·6
		Total	23·2

most useful type of voltage controlled reactance, the criterion of constant loop gain is satisfied best by using an oscillator which is tuned by a variable inductor with the v.c.d. connected in parallel across the tuned circuit. This ideal arrangement is not often used in practice, mainly because it is difficult to make a variable inductor with adequate Q at frequencies below about 10Mc/s. It is therefore necessary to make allowances for the variations of loop gain with frequency which occur when alternative systems are used. The purely electronic system described above is an extreme example of what may happen. In the uncompensated case the loop gain can vary, under adverse conditions, by as much as 40 to 1 over a half-octave band. Thus it is necessary to include a non-linear d.c. amplifier in the control loop to compensate for the non-linear transfer characteristic of the slave oscillator. By this technique gain variations can quite easily be reduced to less than 4 to 1.

Fig. 4. Phase loop response characteristics



When the oscillator is coarse tuned by a variable capacitor the loop gain variation can usually be accommodated providing the tuning inductor is band-switched at intervals of less than one octave. The acceptable variation of loop gain is determined by the capture range required, which in turn determines the absolute value of loop gain. A capture range of 5 per cent has been found to be compatible with a loop gain variation of not more than 5 to 1.

The cut-off frequency and attenuation slope of the phase loop low-pass filter can also affect loop stability. When the capture range is less than 5 per cent the filter cut-off frequency is considerably higher than the bandwidth of the control loop and its effect on stability is small (see Fig. 4(a)). For the electronic tuning system with its hold-in range of half an octave it is necessary to modify the loop response with a lead-lag network[5] shown in Fig. 4(b).

Another important aspect of the tuning system is the short term stability of the slave oscillator. Because the control bandwidth is small no feedback is available to correct short term variations of oscillator frequency at rates above about 1c/s. For this reason it is important to choose an oscillator which is insensitive to vibrational disturbance. Again this is especially true for oscillators which are to be used in mobile s.s.b. equipments. Traditional air-spaced gang capacitors are unsatisfactory in this respect and if shaft tuned circuits are essential a solid dielectric variable capacitor or a rigid inductance-tuner must be used. By these means it is possible to make oscillators whose frequency is deviated by less than ± 5c/s when vibrated at 100c/s with an acceleration of $1g$.

Frequency Selection

The output frequency of the controlled oscillator is determined by information fed into the variable ratio divider. A number of wires, each carrying direct voltages supplies this information in the form of a coded pattern of binary digits which corresponds to the frequency required. The code pattern can be produced either by using coded switches for each digit or by using a diode matrix in association with a conventional single pole multi-way switch. The coded switch uses cams to close up to five contacts on each digit switch and the code pattern is predetermined by the contour and orientation of the cams. The coded switch or diode matrix can cope with any code pattern that could conceivably be required. However in some circumstances the code pattern required for each digit is cyclic and one can then use a simpler type of conventional switch which requires no diode matrix.

Applications and Performance

The performance of a conventional component version of the synthesizer has already been evaluated in field use, and several versions constructed from integrated circuits are currently being evaluated in the laboratory.

The conventional version mentioned above is an interesting example of the versatile nature of the synthesizer. In this case a common synthesizer can function as a frequency generator for either of two completely independent radio equipments. The first is a v.h.f., f.m. radio relay wireless station (including a separate transmitter and receiver) which operates in the band 50 to 100Mc/s with a channel spacing of 125kc/s. The second is also a radio relay station; this time a u.h.f., f.m. equipment operating in the band 225 to 400Mc/s with a channel spacing of 500kc/s. These radio sets were each originally designed for direct control of frequency by a single crystal, the crystal being changed every time a new channel was required. The new synthesizer is used as a direct replacement for the crystal and its output plugs into the crystal socket. To operate a complete radio relay station two synthesizers are required since the receiver and transmitter operate simultaneously but on different frequencies. However each synthesizer can perform the transmit or receive role for both the v.h.f. and the u.h.f. stations. Thus maintenance is simplified and only four crystals are now required to supply all the available channels instead of the original 750. Each equipment contains crystal frequency multipliers and heterodyne mixers. The relationship between aerial frequency and crystal frequency is there-

TABLE 4
Performance of Digital Frequency Synthesizers

CHARACTERISTIC	MAN-PACK H.F., S.S.B.	VEHICLE V.H.F., F.M.	AIRBORNE V.H.F./ U.H.F./A.M.
Minimum output frequency (Mc/s)	3·750	39·6875	150·000
Maximum output frequency (Mc/s)	31·749	109·6625	349·975
I.F. sidestep frequency (Mc/s)	+1·750	+9·6875	+ 50·000
Channel increments (kc/s)	1	12·5	25
Unwanted outputs			
(a) spurious A.M.* (%)	0·2	—	0·2
(b) spurious F.M.* (c/s pk)	10	30	100
(c) non-harmonic † single frequency(dB)	−80	−130	−80
Power consumption** (w)	1·0	1·5	2·5
Volume** (in³)	10	10	15

* *For modulating frequencies in the band 300 to 3 000c/s*

† *Amplitude of single frequency components relative to carrier amplitude*

** *These figures refer to the frequency control unit and do not include either the slave or the standard oscillators*

fore complex and varies both between transmit and receive modes and with the band of frequencies in use. By using the switchable 'sidestep' property of the variable ratio counter all these variations can be catered for in a single synthesizer. A single slave oscillator covers the frequency range 24 to 52Mc/s and feeds the u.h.f. sets directly. The first sub-harmonic, obtained from a binary counter, supplies the v.h.f. sets in the band 12 to 26Mc/s. Apart from some additional buffer amplifiers and simple fixed ratio counters the synthesizer is no more complex than a normal synthesizer which performs only a single role.

In Table 4 the main characteristics and the measured performance of three versions of the digital synthesizer are recorded. Where the performance quoted is well in advance of that commonly achieved by more familiar types of synthesizer, it is reasonable that the reader should query the measurement techniques used. In particular, the figures given for purity of the output signal warrant some explanation of the way in which extremely small spurious outputs are measured.

Two pieces of test equipment have been devised; one to measure unwanted f.m. of the carrier and the other to measure the relative amplitude of non-harmonic, single frequency spurious components of the output. The f.m. detector consists of a multiple heterodyne frequency translator whose final fixed frequency output is at about 15kc/s. F.M. of the output signal is converted to a.m. by applying the output to the attenuation slope of a low-pass filter. The envelope of the 15kc/s signal is observed on a high gain low frequency oscilloscope. Deviations as small as 1c/s, at rates between d.c. and 300c/s, can be detected in this way, without using d.c. amplifiers which suffer from drift and semiconductor $1/f$ noise. The translator uses crystal controlled local oscillators and can operate at a number of spot frequencies within the required band. The residual f.m. of the measuring equipment has been checked using a highly stable crystal controlled source and was found to be less than 1c/s. The equipment is easily calibrated using a l.f. decade oscillator and a frequency counter.

A specially developed sensitive receiver has also been constructed to measure single frequency spurious outputs. This receiver has sufficient dynamic range for it to be able to detect a 1μV spurious signal in the presence of a carrier e.m.f. of 0·5V r.m.s. The adjacent channel selectivity and blocking characteristics of the receiver are such that this sensitivity can be maintained when the wanted and unwanted outputs are separated by only 25kc/s. A single superheterodyne arrangement with an extremely low intermediate frequency is used to make a receiver with very few spurious responses. Since the receiver can be operated with its local oscillator either above or below the input signal frequency such spurious responses as do exist, can be recognized and avoided.

Conclusions

The foregoing description of the digital frequency synthesizer has shown how it can be adapted for a wide range of applications in mobile radio equipment. Because the synthesizer relies on closed loop control of a single slave oscillator, the radio set designer can continue to use the traditional v.f.o. as a transmitter master oscillator or receiver local oscillator. The digital synthesizer provides a precision frequency control system for one or more v.f.o.'s and these oscillators can be associated, in the radio set, with other r.f. circuits to form a ganged tuner assembly.

The development of the oscillator and the frequency control system can proceed independently providing that the inter-face parameters are accurately defined and the two design teams co-operate fully.

The use of common modules, composed mainly of assemblies of digital semiconductor integrated circuits, has simplified the development of a range of synthesizers for mobile applications where small size, low power consumption and high reliability are all important. A consumption of 1W and a volume of 10in³ have been achieved for a precision frequency control unit for use in man carried portable radio sets.

In vehicle borne installations which call for high power transmitters and sensitive receivers to operate simultaneously and in close proximity, the synthesizer must provide an extremely pure output in order to avoid mutual interference. For this role synthesizers with a spectral purity approaching 130dB are being developed. The possibility of achieving this purity has already been demonstrated in the laboratory, where considerable precautions have been taken to isolate the slave v.f.o. from the frequency control system.

The versatility of the digital synthesizer results from the exclusive use of aperiodic circuits throughout. The techniques described can be applied in the large majority of known applications for synthesizers in communications systems with all the advantages already described.

Acknowledgments

Methods of digital frequency synthesis were investigated, in the first instance, as part of a Plessey-UK Ltd research programme. However, much of the work directed at the particular equipment described in this article was carried out in connexion with Ministry of Aviation contracts. The author wishes to thank the Ministry of Aviation and the Directors of the Plessey Co. Ltd, for permission to publish this article.

Digital frequency synthesizers are being developed by a team of engineers working in the Electronics Research Laboratory and the Telecommunication Division of Plessey-UK Ltd, West Leigh. The author gratefully acknowledges the contributions of members of the teams and of the design authority, Signals Research and Development Establishment, Christchurch.

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Point on a Cycle Switch

By R. A. Stevens*, M.Sc., M.I.E.E.

In many instances when investigating transient conditions in a.c. circuits it is necessary to be able to switch on at a predetermined point on the supply wave and to repeat the switching point accurately. Apparatus for carrying this out has been made before but it was decided in this instance to produce a completely solid state system. It makes use of a fairly well established phase shifting technique involving a gate which operates at a selected level on a voltage ramp. The unit is compact and portable and is very useful in undergraduate laboratory work, especially as it has a repetitive mode of operation.

(Voir page 345 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 352)

THE requirement which led to this development was for a switch which would close at any desired point on the cycle of a 50c/s a.c. supply. The phase α of a point of closure was to be variable between 0 and π with an accuracy of $\pm 2°$ as illustrated in Fig. 1(a). It will be seen that after the first half cycle the switch is to remain closed for the whole of each subsequent half cycle unlike the usual welding control where closure is delayed in each half cycle as shown in Fig. 1(b).

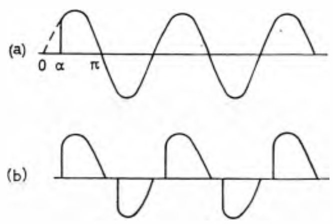
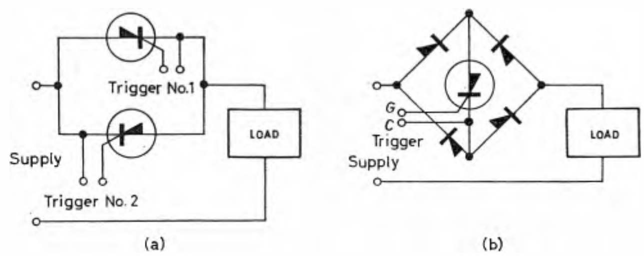


Fig. 1 (left)
(a) Waveform produced by switch
(b) Waveform produced by normal thyristor contactor as used for welding control

Fig. 2 (right)
(a) Double thyristor switching circuit
(b) Single thyristor and diode bridge circuit



The switch is to be used partly as a teaching aid in normal undergraduate laboratory experiments and partly in connexion with investigations on transients in power systems.

This type of switching has been carried out in several different ways in the past using synchronous electromechanical contactors, synchronous rotary switches, thyratrons, ignitrons or transducers[1,2,3].

It was decided in this instance to use a thyristor as the switching element and to use solid state devices in the control circuits. Also it was decided that if possible a single 10 volt power supply would be used. This precluded the use of two thyristors back to back, as d.c. triggering of the thyristor was required for reasons of accuracy of the phase of the switching point and would require two entirely separate electrically isolated control circuits, Fig 2(a). Hence a silicon diode bridge was used with the thyristor connected diagonally as shown in Fig. 2(b).

Normally the device would be used with some form of recording oscillograph such as an ultra-violet recorder, single shot recordings being made. It was desired, however, to incorporate a repetitive mode in which the switching would be synchronized with the time-base of a c.r.o. In this mode, after closure several cycles would appear on the c.r.o., the switch would then open and after a suitable interval to allow the circuit under test to settle, the time-base of the c.r.o. would be triggered and immediately afterwards the switch closed again. Thus a stationary pattern would appear on the c.r.o. It is advantageous to use a long persistence tube.

Switching off will always occur at the end of the half cycle in which the gate current of the thyristor is reduced to zero, the current will thus always naturally fall to zero at the end of this half cycle. If then the switch is used to examine switching transients in an iron cored inductor or transformer, the core flux will always return to the remanent value.

This raises certain difficulties when different initial values of core flux are required, but these can be overcome by the use of certain auxiliary circuits which are not discussed here.

The switch is made in two parts with the control circuits as one unit which will deliver a maximum of about 300mA to trigger a thyristor, and a second unit which is the thyristor bridge. This enables a control unit to be used with any thyristor bridge whose rating may be chosen to suit any particular situation which may arise. The output of 300mA is suitable for triggering most sizes of thyristors[5].

Circuits

The circuit is based on a thyristor firing circuit in which the instant of conduction is determined by the time taken for a ramp voltage to rise to a set value which can be varied to alter the firing point.

The system can be broken down into six sections, as shown in Fig. 3, which are described in turn and are:

(1) Pulse generator to produce 50c/s timing pulses.
(2) Ramp generator and gate.
(3) Thyristor drive amplifier.
(4) Logic circuit.
(5) Astable multivibrator for repetitive working.
(6) Thyristor bridge.

The circuit details are shown in Fig. 4.

Sections (1), (2) and (3) are based on a circuit which is fully described elsewhere[4] so that only a brief description of the operation, and no design details will be given here.

(1) MR_1 conducts on the negative half cycle of the a.c. input voltage, cutting off VT_1. On the positive half cycle, base current flows via R_1 and the collector potential of VT_1 falls, Fig. 5(a) and (b). This is differentiated by C_1 R_3 giving alternate positive going and negative going pulses at the base of VT_2. VT_2 will conduct on the positive going pulses, hence negative going pulses appear at its collector, Fig. 5(c) and (d).

(2) VT_3 conducts on the negative going pulses and dis-

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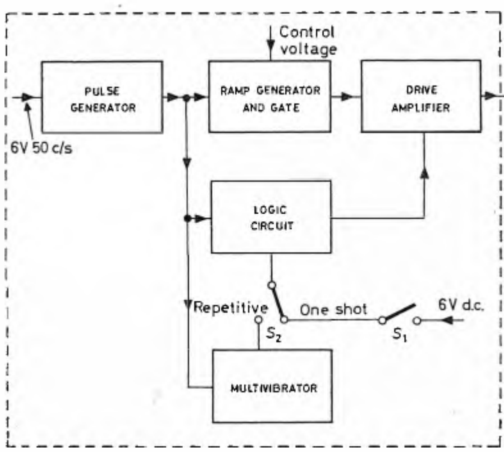


Fig. 3. Arrangement of the complete switching scheme.
The control unit itself is contained within the dashed line

charges C_2 which will commence to charge via R_9 at the end of the pulse A in Fig. 5(d). C_2 will charge until the emitter potential of VT_4 rises above its base potential, point B in Fig. 5(e), when it will conduct. The base potential of VT_4 is controlled by the base potential of VT_5, thus the setting of RV_1 will decide the instant at which VT_4 conducts. VT_4 is a silicon transistor so that the leakage current variation is small.

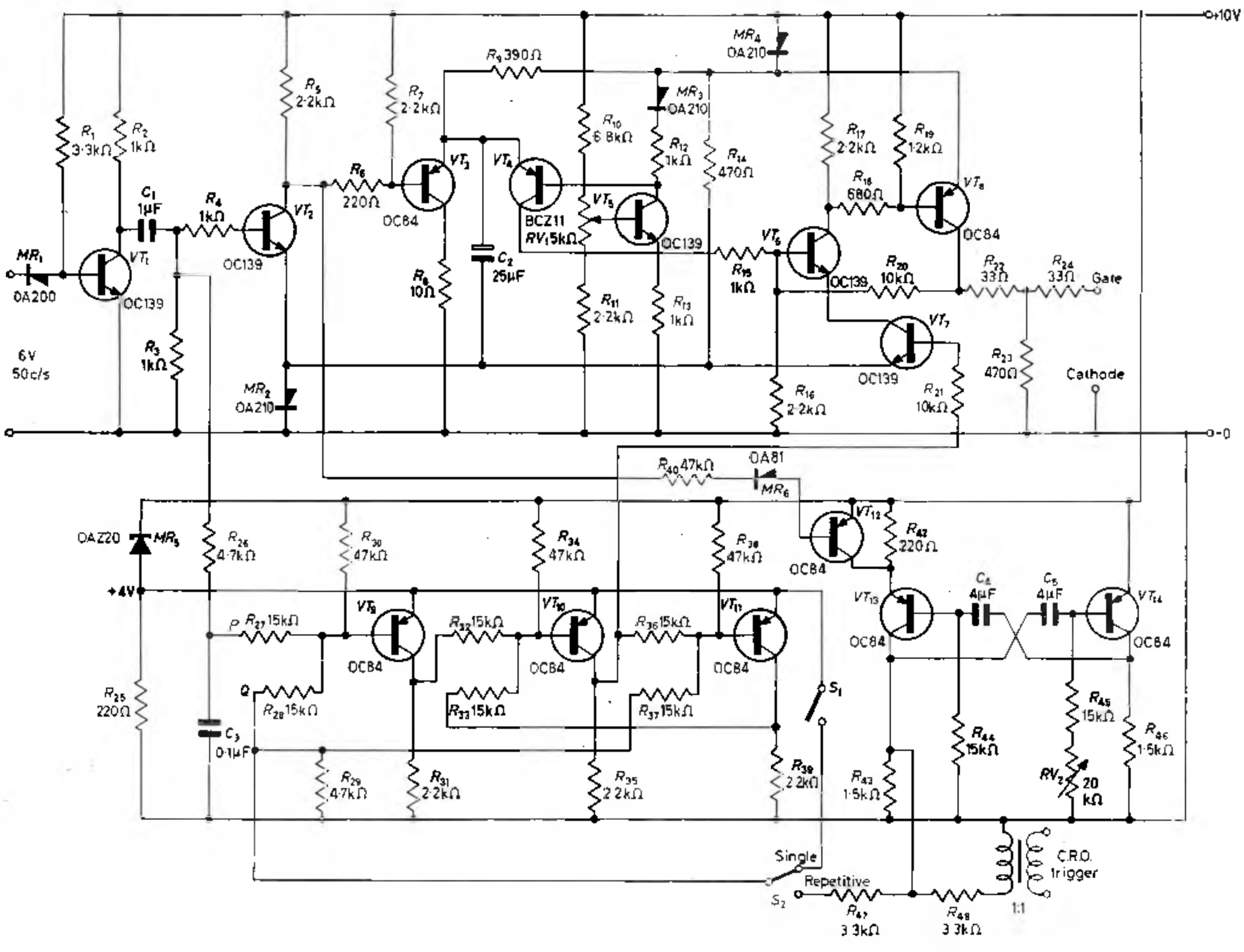
(3) When VT_4 conducts, the base of VT_6 becomes more positive than its emitter and, provided VT_7 is in a conducting condition, it will 'bottom'. In turn VT_8 'bottoms' and a positive voltage appears at its collector, Fig. 5(f). This triggers the thyristor and causes it to conduct. In addition a positive voltage is fed back to the base of VT_6 via R_{20} which holds VT_6 and VT_8 'on' and ensures that the thyristor conducts over the whole of the succeeding half cycles. Fig. 5(g).

(4) If the master switch S_1 simply applied a positive voltage to the base of VT_7 the time of triggering the thyristor would be decided by either (a) the conduction of VT_4 if this occurred after the closure of S_1 or (b) by the closure of S_1 if this occurred after VT_4 started to conduct. The latter possibility must be eliminated. If S_1 is closed after VT_4 commences to conduct in any one cycle, the thyristor must not be triggered until after VT_4 conducts on the next cycle, Fig. 6.

This is assured by the use of VT_9, VT_{10} and VT_{11} which are used as NAND gates. VT_9 will be conducting unless both points P and Q are made positive. When S_1 is closed VT_9 will cut off only when a positive pulse arrives at P from the base of VT_2. This will switch VT_{10} on, and as the positive voltage from S_1 is also applied to VT_{11}, VT_{11} will cut off. The feedback from the collector of VT_{11} will ensure that VT_{10} remains 'on'. The positive voltage from the collector of VT_{10} will switch VT_7 on and enable the thyristor to trigger as soon as VT_4 conducts.

This ensures that triggering cannot take place merely

Fig. 4. The control unit



on closing S_1, but only after the next time that the timing circuit C_2 R_9 has been reset.

The delay network R_{26} C_3 was found to be necessary as there was a tendency for VT_7 to switch on slightly before VT_4 had cut off at the end of its conducting period thus triggering the thyristor too early.

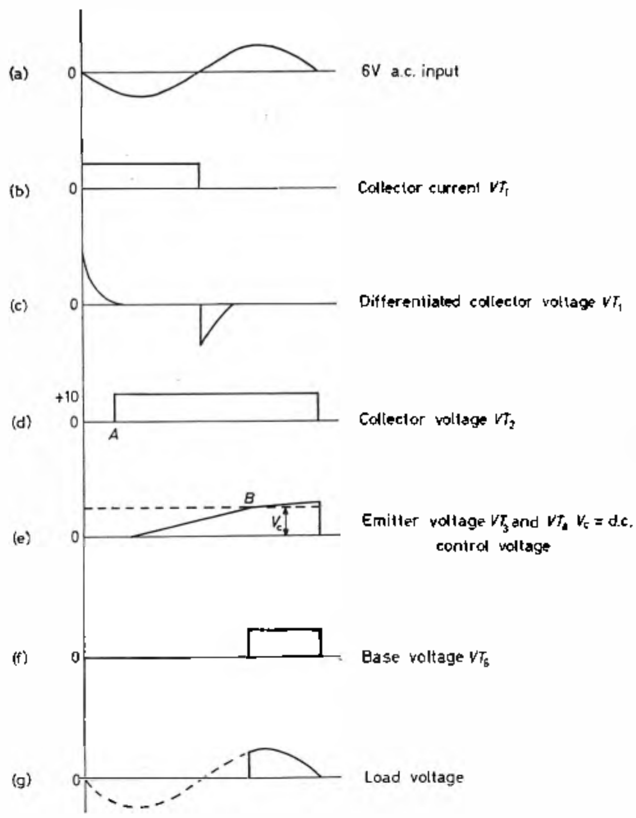
(5) VT_{13} and VT_{14} form an assymetrical multivibrator in which the 'on' time of VT_{13} can be varied by means of RV_2 from about four to seven cycles of the a.c. supply. The instant at which VT_{13} is switched on is synchronized to the a.c. supply by means of pulses from the collector of VT_2 fed to the base of VT_{12} via R_{40} and MR_6. The 'on' time of VT_{14} is three cycles of the supply.

In the repetitive mode the voltage at the collector of VT_{13} is taken to the input Q of the logic circuit thus triggering the thyristor repetitively at a rate decided by the multivibrator frequency. The voltage at the collector of VT_{13} is also taken via R_{48} to the primary of an isolating transformer to provide negatively going pulses of about 4V amplitude to trigger the cathode-ray oscilloscope. Isolation is necessary as there is no direct electrical connexion between the negative line of the control unit and the a.c. supply which is being switched and to which the c.r.o. will be connected.

(6) As has been stated previously the bridge circuit was chosen so that only one thyristor has to be triggered. If pulsing is permitted, a single control unit can be used with back-to-back thyristors as there can be an output transformer with two secondary windings one for each thyristor, but it was considered that d.c. triggering would give a more precise control of phase.

It must be borne in mind when considering ratings that the thyristor will carry current on both cycles, whereas in the two thyristor contact circuit each carries current for only half of each cycle. Hence a larger thyristor will be required in addition to the four diodes in the bridge, however at the present time the bridge discussed here will probably cost less than the alternative. R_{24} is included to give some control of the gate current and should be chosen so that the current has the minimum value which will give consistent firing for the thyristor which is to be used, this avoids excessive gate dissipation particularly with smaller sized thyristors. The value shown is for use with the Mullard BTY87-400R which is a 12A thyristor with a 400V p.i.v. rating. The diodes used are Mullard type BYZ12.

Fig. 5. Waveforms at the various points in the pulse generator, ramp generator, gate and drive amplifier circuits



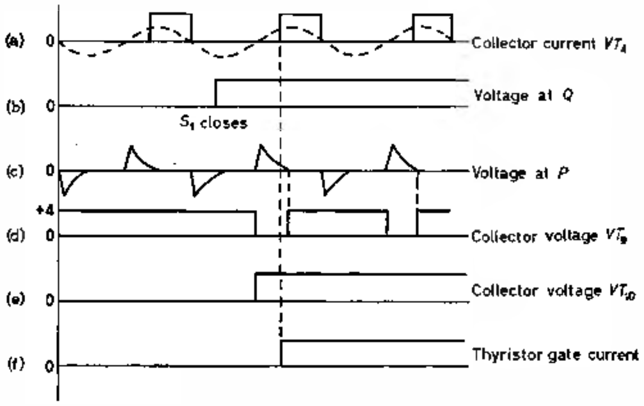


Fig. 6 (above). Waveforms at various points in the logic circuit

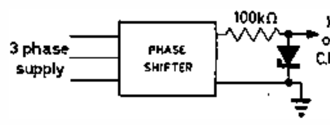


Fig. 7 (right). Circuit used to produce a square wave of calibration phase for calibration purposes

Calibration and Accuracy

Calibration was carried out using the circuit of Fig. 7. The Zener diode produces a rectangular waveform which can be shifted in phase by the phase-shifting transformer. This is applied to one beam and the switched waveform to the second beam of a double beam c.r.o., thus the phase delay can be measured by shifting the phase of the comparison waveform into coincidence with the switching point. It is found that the process is much easier and more accurate using the rectangular wave than it is with a sine wave.

The phase was found to vary not more than $\pm 2°$ with a ± 5 per cent variation of the 10V d.c. supply and the calibration held within these limits from day to day. Variations in the values of R_9 and C_2 will affect calibration, but it would be possible to alter their values, keeping the product about the same, so that a paper or polystyrene dielectric capacitor could be used. R_9 should not be too large however as the leakage current of VT_3 may then be significant.

Tests have been carried out using the apparatus to examine the waveform of the magnetizing current surge on switching a transformer on to an a.c. supply with satisfactory results.

It is hoped to extend the scope of the equipment later for use on three-phase supplies.

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A Period Modulation/Demodulation System

By L. Molyneux*, Ph.D.

A circuit is given and the performance demonstrated of a period modulator/demodulator system. The main advantage of the system over frequency modulation, which it closely resembles, lies in the simplicity of the low-pass filter following the demodulator, in some cases the filter can be omitted altogether.

(Voir page 345 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 352)

PERIOD modulation[1], where the time between similar parts of the carrier waveform is made a linear function of the modulating voltage, can be demodulated so that the energy at carrier frequency (and its harmonics) at the demodulator output is proportional to the change in the modulating voltage in the period between successive cycles of the carrier. This is in contrast to frequency modulation, which it otherwise closely resembles, where the unwanted energy is largely independent of the modulation. The filter following a frequency demodulator, particularly where the modulation frequency approaches the mean carrier frequency within a factor of 10 or less (as often occurs in tape recording installations) must have a flat characteristic up to the highest modulation frequency, and then an abrupt change of attenuation in order to suppress adequately the carrier. This high attenuation at the carrier frequency is needed since even at low or zero modulation the components of the carrier have their full value and are therefore most troublesome. Such filters cannot be constructed out of resistors and capacitors alone and inductors are needed. It should be said, however, that circuits can be made with the required characteristics using only resistors and capacitors if they also incorporate some active circuit components[2]. These circuits are not yet common but development work on them is being actively pursued.

In the case of the period demodulator, components of the carrier and its harmonics are in theory completely absent at zero modulation and thereafter increase in proportion to the modulation, the filter following the demodulator can therefore be of simple resistance-capacitance design or even omitted altogether. The complete modulator/demodulator system can be built without inductors and it may be realized in integrated or thin-film circuit form.

Modulation

The essential components of the modulator are: a linear ramp generator and a voltage comparator. The ramp and input voltage, biased if necessary so that the resultant is always of one sign, are connected to the voltage comparator. When the output from the voltage comparator indicates equality of its inputs the ramp is returned to zero in a time short (~ 10μsec) compared with the mean carrier period (~ 1msec) and a new ramp started. Each time the ramp is returned to zero an output pulse is formed and this pulse train, possibly modified in shape, forms the output of the period modulator. As the ramp rises at a constant rate (Fig. 1) the time between successive output pulses is a linear function of the absolute value of biased input voltage and period modulation is achieved.

A practical circuit is shown in Fig. 2. RV_1 sets the bias voltage at a suitable value (~$\frac{1}{2}$V supply). The input is floating with respect to the neutral line and is effectively in series with the bias and the input impedance convertor, VT_1 and VT_2. VT_6 is a constant current generator, the current being defined by the potential divider connected to its base and the value of the emitter resistor. The constant current from VT_6 causes a negative-going ramp to develop across C_1, when the ramp is less negative than

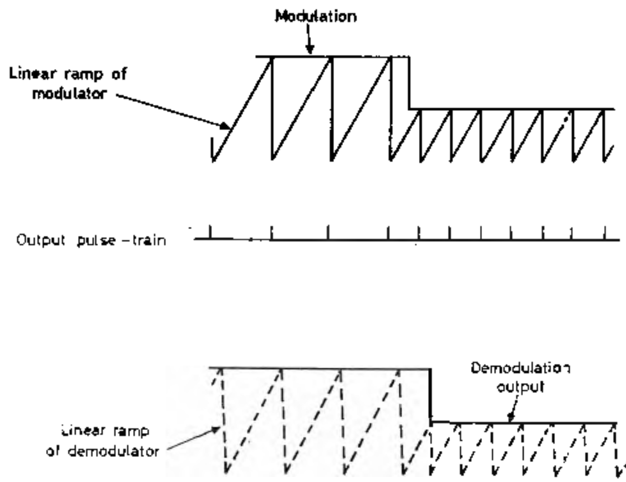


Fig. 1. Waveforms of modulator and demodulator

the biased input voltage, VT_3 draws no base current; base current flows when the ramp is slightly more negative than the base of VT_1 and the consequent collector current is amplified by VT_4. The comparator therefore consists of VT_1 VT_3 and VT_4; VT_1 serving in a double capacity of part comparator, part impedance convertor.

The output from the comparator triggers a complementary bistable VT_7 and VT_8 into the conducting state when the collector of VT_8 goes negative causing VT_5 to conduct and discharge C_1 towards the neutral line. When C_1 is within a volt or so of the neutral line MR_1 diverts the base current of VT_7 and turns the bistable off and with it VT_5, allowing a new ramp to develop. The circuit in Fig. 2 is intended for tape recording but the narrow (~ 10μsec) pulse at the collector of VT_8 is unsuitable as a recording signal, it is therefore used to trigger a monostable VT_9, VT_{10} which produces a pulse of about 0·5msec duration.

Demodulation

The essential features of the demodulator are: a ramp generator similar to that in the modulator, a means of transferring the instantaneous value of the ramp to a storage capacitor and a means of returning the ramp to zero.

When an input pulse is applied to the demodulator a storage capacitor is immediately (~10μsec) set to the value of the ramp at this instant then, after a short (~ 2μsec)

* *The University of Newcastle upon Tyne*

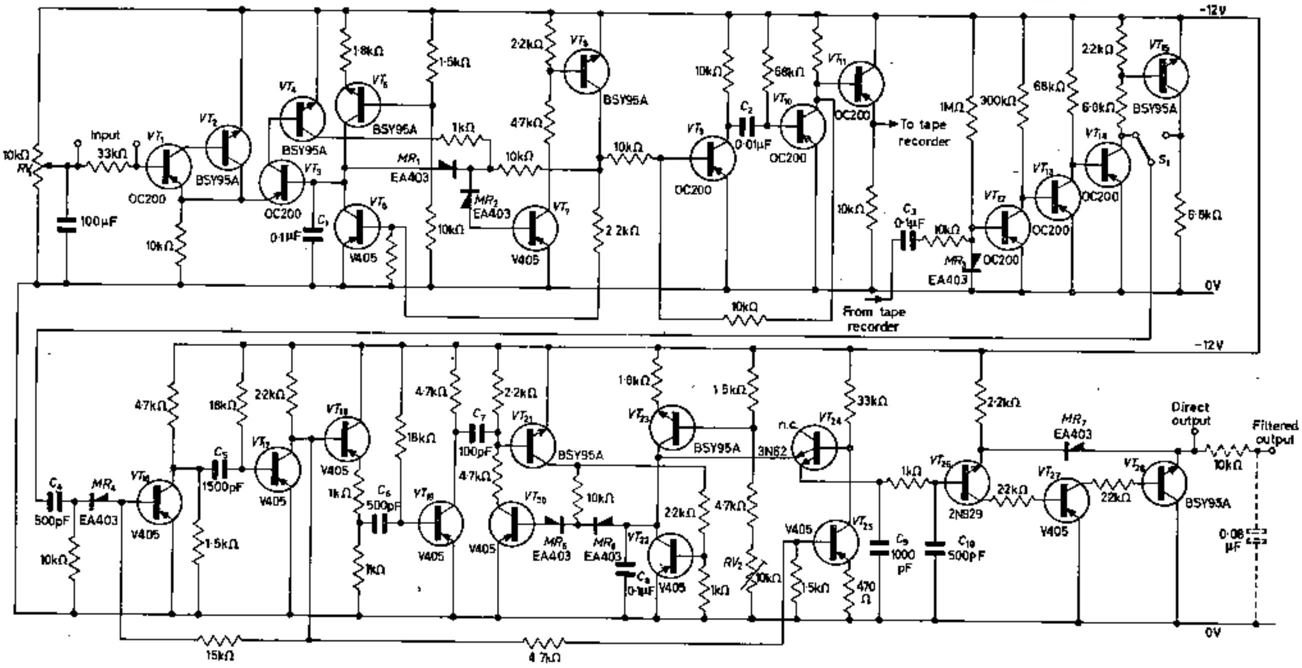


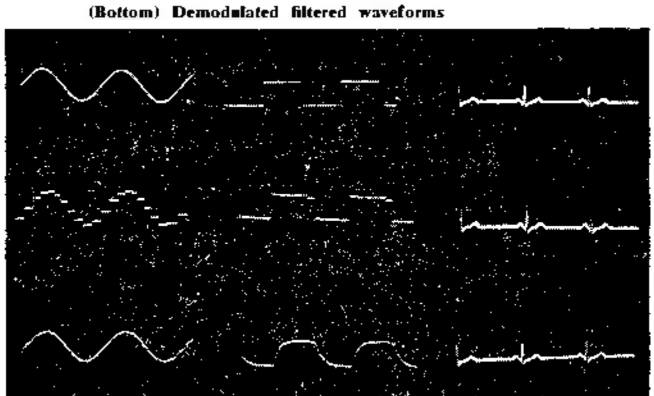
Fig. 2. Practical circuit of a period modulator and demodulator
(NOTE: *Base-emitter resistor of* VT_5 = *1k*Ω; *collector resistor of* VT_{10} = *10k*Ω)

delay, the ramp is returned to zero in a manner similar to that of the modulator.

The practical circuit starts with a section necessary to deal with the distortion of the tape-recorded signal. The bandwidth of a typical tape recorder with slow tape speed ($1\frac{7}{8}$in/sec) is too small ($\sim$ 5kc/s) to allow accurate reproduction of the input pulse, and the output waveform (magnitude $\sim$ 2 volts*) although having the same period as the input has lost its sharp edges. It is therefore amplified by VT_{12}, VT_{13} and VT_{14}, each stage, including the first, being driven far beyond its linear range with the result that the output of VT_{14} is of almost rectangular form. However as the input to VT_{12} is not symmetrical one edge is probably better defined than the other and S_1 is used to select the best edge to trigger the monostable VT_{16} and VT_{17}. When the monostable is triggered the collector of VT_{17} goes negative, causing VT_{25} to draw current. The current is limited by the emitter resistor to about 2mA and most of this current is available to drive the base of a double emitter transistor VT_{24}, bringing the two emitters to the same potential and setting C_9 to the value of C_8. C_8 has the ramp developed across it and therefore C_9 is set to the value of the ramp at the time of the input pulse. VT_{24} used with an open-circuited collector behaves as a symmetrical transistor with a current gain of about 2, therefore a charging current of about 4mA is available and this is more than sufficient to change the potential of C_9 and C_{10} by 2V, which represents a typical maximum modulation excursion, as would be found for instance, if a square wave of $\pm$1V were being applied to the modulator. At the end of the 10μsec transfer pulse VT_{19} and its associated resistor and feed capacitor introduce a 2μsec time delay before initiating the ramp reset signal which triggers VT_{20} and VT_{21} and resets the ramp voltage by means of VT_{22}. This delay allows the transfer switch (VT_{24}) to turn completely off before returning the ramp to zero. In a manner similar to the modulator a new ramp begins to form immediately after the last one has been returned to zero. The slope of the demodulator ramp can be adjusted to 'match' that of the modulator by RV_1.

VT_{26}, VT_{27} and VT_{28} form a high impedance convertor so that the charge on C_9 does not leak away appreciably between input pulses while MR_7 ensures that the output voltage will be at the same level as the input to the base of VT_{26}. The resistor and capacitor in the base lead of VT_{26} limit the rise-time of a step change to a value that lies within the pass-band of the three transistors VT_{26}, VT_{27} and VT_{28} and prevents transient disturbances of the output. The output for zero modulation (constant ramp height) is a constant voltage and the output for a varying voltage at the modulator is a stepped waveform, each step having a height equal to the difference in modulation between pulses of the carrier wave. This is illustrated in Fig. 3 where sine and square waves at 100c/s are used to modulate a 1msec period carrier wave (1 000c/s) at a modulation depth of $\pm$0·2msec. The top row shows the modulating waveforms. The middle row

Fig. 3. (Top) Modulating waveforms (100c/s sine and square-waves and 1c/s e.c.g. type waveform)
(Middle) Demodulated unfiltered waveforms
(Bottom) Demodulated filtered waveforms



* *Tandberg Ltd Model 62.*

shows the output of the demodulator while the bottom row shows the effect of a single stage resistance capacitance filter (3dB point 200c/s). The extremely good transient response of the circuit to the square-wave in the unfiltered case is not all it seems to be. The fast ($\sim 10\mu$sec) edge of the output is generated by the demodulator and of itself implies a broader bandwidth than the system possesses. However, it is independent of the input waveform rise time if this is shorter than a period of the carrier wave and the time at which it takes place cannot be defined to better than one carrier period so the effective bandwidth is no better than the transmission system.

The circuit is particularly suitable for recording physiological waveforms such as are produced by the heart (e.c.g.) and brain (e.e.g.). For these waveforms 100c/s is an adequate bandwidth, and they can be successfully recorded at 1⅞in/sec. The last column (Fig. 3) shows an e.c.g. type waveform (actually from a generator[3]) with a repetition rate of about one per second. The modulating waveform is, as before, at the top; the middle and bottom waveforms are without and with filtration respectively. It can be seen that in this case filtration man be dispensed with. It is an interesting and useful feature of the system that if the ramp waveforms of the modulator and demodulator have the same slope then the output from the modulator has the same magnitude as the potential applied to the voltage comparator and the system has an inherent overall voltage gain of unity. Further, if the ramp generators in the modulator and demodulator have the same waveforms the overall transfer function of the system will be linear even if the waveforms themselves are not.

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Thyristor Control of D.C. Motors

Four thyristor phase control systems to drive a shunt wound d.c. motor have been designed by the Industrial Market Division of Mullard Ltd.

The method consists of adjusting the conduction of 'triggering' angle of the thyristors so that the amount of power supplied to the motor from a single or three phase supply can be varied, thus controlling the motor speed.

These control systems range from a simple open-loop system with manual control of speed to a complex system with closed control of speed, overload protection and provision for speed control from an external isolated control voltage.

Additional facilities can be added to any of the systems such as current limiting, controlled acceleration, automatic protection against loss of motor field, compensation for mains voltage variations and motor field loadings.

Remote control of speed can also be added since the thyristor system can be operated by a remote pick-up, a feature which is an advantage in automatic production lines.

A brief description of the four systems is as follows:

Simple Speed Control Circuit

In this system, speed control is based on the measurement of the back e.m.f. of the motor and using this in a feedback circuit to stabilize the speed. A current-limiting circuit is included to prevent the motor current rising appreciably if the motor is overloaded. This circuit also limits the triggering current during acceleration when starting by retarding the thyristor angle so that the conduction angle is small. The circuit is designed so that, on switching on with the speed control set to maximum speed, the input current never exceeds the current limit value, even during the first few half-cycles of the supply.

In the main speed-control feedback loop, a signal is taken from an armature back-e.m.f. bridge and fed to a thyristor trigger module. Here it is compared with a reference voltage provided by a voltage regulator diode and used to control the thyristor triggering angle, in order to correct the changes in motor speed brought about by changes in the load torque. An overload clamp forms part of the back-e.m.f. bridge, and is adjusted so that the part of the feedback signal, which is proportional to the armature current, is removed when the current reaches the maximum rated value. The clamp serves two purposes. It acts as a partial current-limit by removing part of the speed-stabilizing feedback; and this assists the true current-limiting circuit which thus requires a lower overall gain. It also modifies the loop-gain characteristic of the speed-stabilizing circuit and increases the stability margin of safety.

Armature-Voltage Speed Control

Here the output from a back-e.m.f. bridge is compared, in a resistive adding network with a d.c. reference signal. Speed is changed by varying the reference voltage. The difference output from the adder is fed to a d.c. amplifier, the output of which is used to control a thyristor trigger module and thence the thyristors in the power control stack.

Additional features in this circuit include a time-constant circuit which prevents the control voltage from being applied to the amplifier input immediately the mains supply is connected, so eliminating starting surges and allowing the current-limiting circuit to take over smoothly during the motor acceleration period.

In addition there is a compensating circuit which takes into account the variation of speed due to changes in field current arising from field temperature changes.

Tachometer Speed Control

The armateure back-e.m.f. bridge circuit of the previous control system is replaced by a d.c. tachometer mechanically coupled to the motor shaft.

Additional circuits are included so that the regulation at all speeds, expressed as a percentage of the actual speed, is nearly constant.

Three-Phase Speed Control

When the armature-voltage and tachometer speed controls are employed the modifications consist of replacing the single phase thyristor circuit by three-phase ones for use of power supply from a three-phase source.

An Optical Communications Link

Where a wired connexion is impractical and a radio link is debarred for reasons of security or severe electrical interference, communication can be established over a line-of-sight path by the use of an optical link. An indefinite number of such links might be used simultaneously without mutual interference and without occupying space in the already-congested radio frequency spectrum.

Apart from other more general applications, there is particular interest in the prospect of using such links for data transmission or voltage and current monitoring in association with extra-high-voltage power lines, where no other method is so convenient for bridging the high potential difference between line and ground.

An equipment developed by Standard Telecommunication Laboratories Ltd, Harlow, Essex, provides a single pulse width modulated channel normally capable of handling a modulation bandwidth extending from 50c/s to 1Mc/s (3dB points).

The light-emitting device used in the present model is an experimental disk diode now under development at STL. The photo-detector is a PCIAF fast silicon photodiode manufactured by STC. A working range of 200ft is achieved with 250mW transmitter drive power.

Bearing in mind the possibility of intense electrical and magnetic interference in a power station environment, the mechanical design of the two heads has been chosen to give a high level of immunity to stray fields. Refracting optics have been used to enable both the source and detector to occupy positions deep within their respective housings. All circuit components, including the power supply batteries, are enclosed in multiple electrostatic and magnetic screens. A single 50Ω coaxial cable to each unit provides the input or output connexion. The batteries enclosed within the transmitter unit are sufficient to provide 16 hours running time; the receiver batteries allow rather more than double this figure.

Critical Inductance for Half-Controlled Thyristor Rectifiers

(Part 2)

By J. L. Storr-Best*, B.Sc.(Eng.), A.M.I.E.E.

(Voir page 273 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 280)

Half-Controlled Three-Phase Bridge Rectifier

Fig. 6(a) is a schematic of a half-controlled three-phase bridge rectifier. As before, only half the bridge arms contain thyristors and a bypass diode is necessary at the output; note that the thyristors could equally well have been placed at the negative end of the bridge in which case their common anode connexion would allow a common heat sink to be used. Fig. 6(b) shows the waveforms and equations of the input voltages, referred to a common neutral. When the firing angle is zero the thyristors conduct almost as soon as their anodes become positive with respect to their cathodes, and the bridge behaves like that of a conventional diode circuit. The output voltage applied to the *LC* filter may be obtained at any point in the cycle, from the waveforms of Fig. 6(b), by subtracting the most negative voltage from the most positive. Thus from 0° to 60° the output is *A-B*, from 60° to 120° it is *A-C*, from 120° to 180° it is *B-C*, and so on. The resultant output waveform is of the familiar 6^{th} harmonic form shown in Fig. 7(a). It will be noted that the first letters in the labels *AB*, *AC*, *BC* etc. indicate which thyristor is conducting and the second letters indicate which diode is conducting.

When firing delay is imposed, the waveform changes in the manner illustrated by Figs. 7(b) to (e). Since only thyristor conduction is delayed, and not diode conduction, successive 60° periods are no longer symmetrical and the periodic frequency changes from the 6^{th} to the 3^{rd} harmonic of the supply frequency[4]. At angles of α greater than 60°, the bypass diode comes into conduction for a portion of the cycle, as indicated by the *D* labels. Full control requires a trigger phase range of 0° to 180°.

The mean d.c. output voltage, for any firing angle, α, may be found by integrating the rectifier output voltage waveform. Despite the variations in waveshape that occur over different ranges of α, it turns out that at any value of α between 0° and 180°

$$V_{dc} = V_{dco} \cdot \frac{1 + \cos\alpha}{2}$$

where V_{dco} is the output voltage at $\alpha = 0°$, and is equal to $3/\pi \cdot E_L$

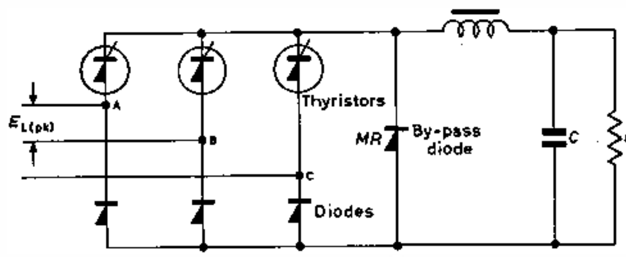
A plot of this relationship between V_{dc} and α may be found elsewhere[5] and is not reproduced here. Three points of similarity to the single-phase case may be mentioned. Firstly the law of voltage reduction with increasing α is the same, and secondly a full range of 180° phase control is required in both cases (and not 120° in the three-phase case as may have been supposed.) Thirdly, the critical inductance is going to remain finite down to zero output voltage since both the d.c. and the a.c. rectifier output voltage components approach zero together.

The shape of the current waveform may be deduced, as in the case of the single-phase rectifiers, by noting the positions of the voltage intersections, where $L\,(di/dt)$ must be zero. Fig. 8(a) shows the waveform intersections and the shape of the current for $\alpha = 0°$. Note that there are four intersections over the 120° period and therefore there are two maxima and two minima.

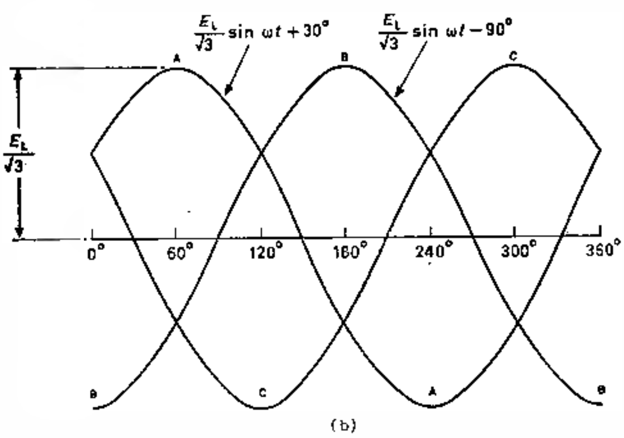
At an angle α of 30° as shown in Fig. 8(b), there are still four intersections but there are now a major maximum and minimum, the middle two being less significant. Going a bit further with α (actually above about 35·5°) the d.c. voltage line fails to intersect the dip in the waveform at the junction of *AB* and *AC* and one is left with only the major maximum and minimum—see for example Fig. 8(c) where $\alpha = 60°$.

Calculating the equation for the instantaneous current, *i*, in the single-phase case, it was found that two ranges of α had to be considered: at angles of α greater than 35·5° the first voltage intersection occurred on the vertically rising edge of the voltage waveform so that $\phi = \alpha$; at lower angles of α, the intersection occurred on the sine wave portion of the waveform, at an angle ϕ greater than α. In the three-phase case, there are the same two ranges to consider for the same reason, but here the changeover

Fig. 6. Half-controlled three-phase bridge rectifier
(a) Circuit arrangement
(b) Voltage waveforms, referred to neutral



(a)



(b)

* *Standard Telephones and Cables Limited.*

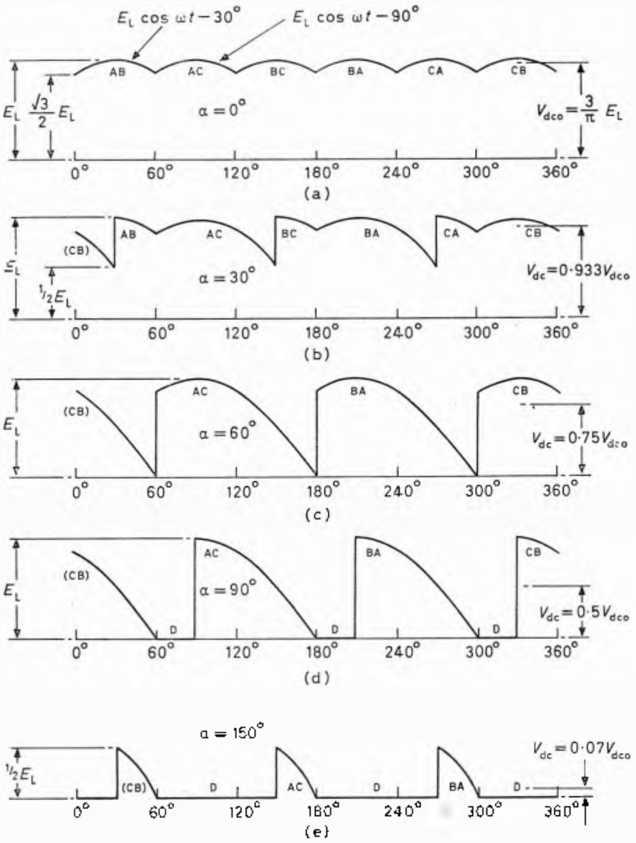


Fig. 7. Half-controlled three-phase bridge rectifier output voltage waveforms at various values of α

occurs where α is about 11° instead of 35·5°; there is a third range in addition to consider for the three-phase case where, at 60°, the waveform changes from a pair of sine wave tops to a single sine wave portion plus a flat portion of zero amplitude (see Figs. 7(b), (c) and (d). There are therefore three distinct ranges of α for which different equations for i, and consequently for $\omega L_c/R$, will apply. These are as follows:

For values of α between 0° and 11°

$$\left(\phi = (\pi/6) + \cos^{-1}\left[3/2\pi\,(1+\cos\alpha)\right]\right)$$

$$i = E_L/\omega L_c \left[\sin(\omega t - 30°) - \sin(\phi - 30°) - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \phi)\right] \text{ from } \omega t = \alpha \text{ to } 60°$$

$$i = E_L/\omega L_c \left[1 - \cos\omega t - \sin(\phi - 30°) - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \phi)\right] \text{ from } \omega t = 60° \text{ to } (120° + \alpha)$$

$$I_{dc} = E_L/\omega L_c \left[\tfrac{1}{2} + (3/2\pi)(\alpha + \sin\alpha) - \sin(\phi - 30°) - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)((\pi/3) + \alpha - \phi)\right]$$

$$\omega L_c/R = \phi - \alpha - (\pi/3) + \frac{\alpha + \sin\alpha + (\pi/3)\left[1 - 2\sin(\phi - 30°)\right]}{1 + \cos\alpha}$$

For values of α between 11° and 60° ($\phi = \alpha$)

$$i = E_L/\omega L_c \left[\sin(\omega t - 30°) - \sin(\alpha - 30°) - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \alpha)\right] \text{ from } \omega t = \alpha \text{ to } 60°$$

$$i = E_L/\omega L_c \left[1 - \cos\omega t - \sin(\alpha - 30°) - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \alpha)\right] \text{ from } \omega t = 60° \text{ to } (120° + \alpha)$$

$$I_{dc} = E_L/\omega L_c \left[\tfrac{1}{2} + (3/2\pi)(\alpha + \sin\alpha) - \sin(\alpha - 30°) - \frac{1+\cos\alpha}{2}\right]$$

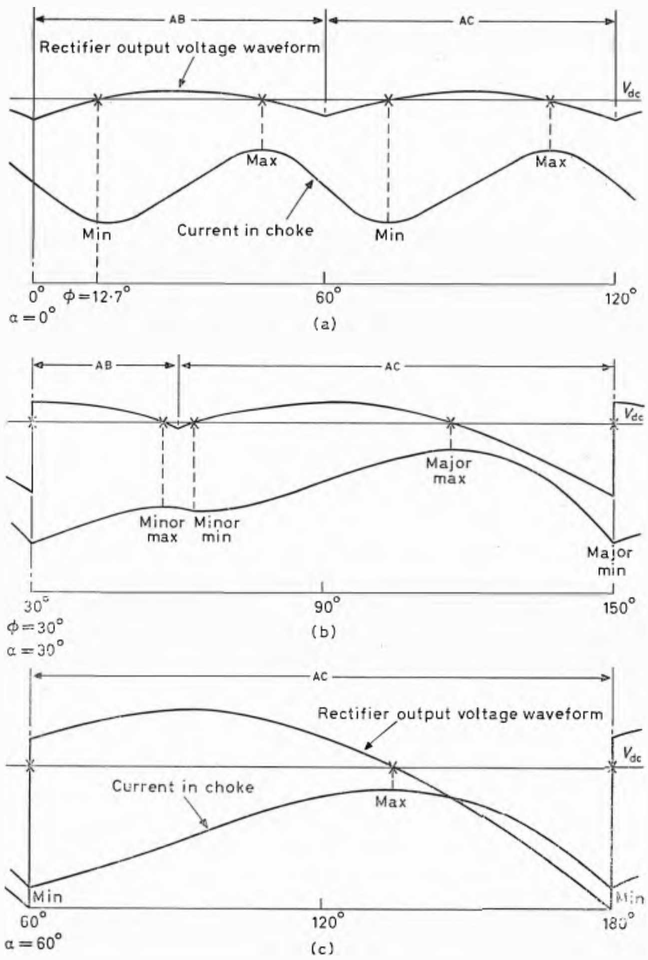
$$\omega L_c/R = -(\pi/3) + \frac{\alpha + \sin\alpha + (\pi/3)\left[1 - 2\sin(\alpha - 30°)\right]}{1 + \cos\alpha}$$

For values of α between 60° and 180° ($\phi = \alpha$)

$$i = E_L/\omega L_c \left[\cos\alpha - \cos\omega t - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \alpha)\right] \text{ from } \omega t = \alpha \text{ to } 180°$$

Fig. 8. Half-controlled three-phase bridge rectifier voltage and current output waveforms for

(a) $\alpha = 0°$
(b) $\alpha = 30°$ (Choke inductance is greater than critical)
(c) $\alpha = 60°$



$$i = E_L/\omega L_c \left[1 + \cos\alpha - (3/\pi)\left(\frac{1+\cos\alpha}{2}\right)(\omega t - \alpha) \right]$$

from $\omega t = 180°$ to $(120° + \alpha)$

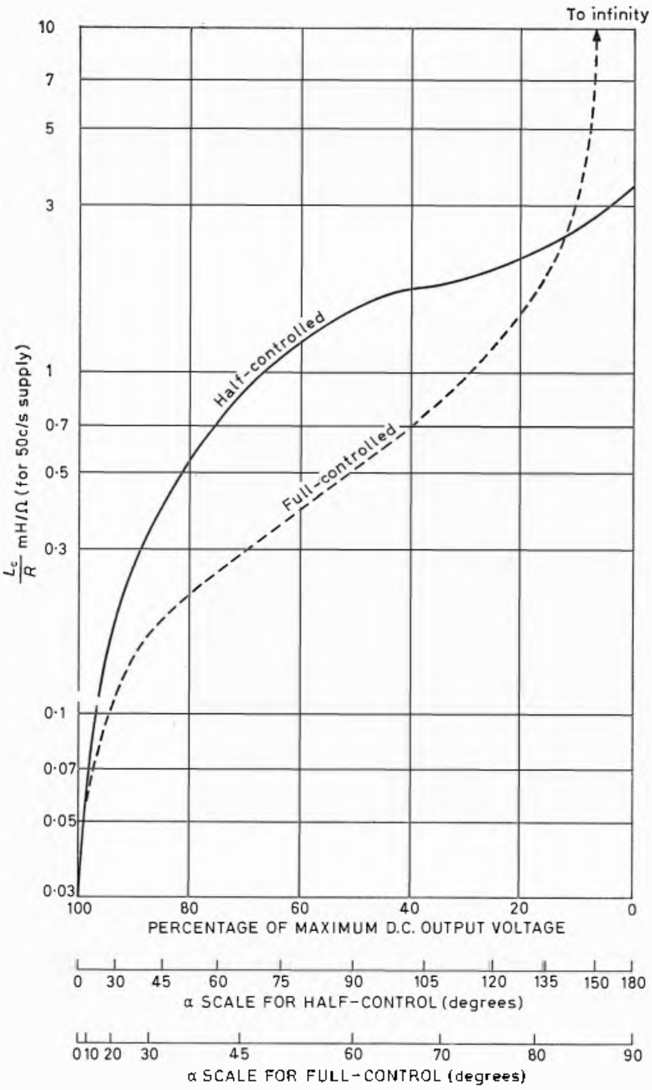
$$I_{dc} = E_L/\omega L_c[3/2\pi\ (\alpha + \sin\alpha) + \tfrac{1}{2}\cos\alpha - 1]$$

$$\omega L_c/R = (\pi/3) + \frac{\alpha + \sin\alpha - \pi}{1 + \cos\alpha}$$

Using these three equations for $\omega L_c/R$ to cover the full range of α from 0° to 180°, the ratio L_c/R has been normalized for 50c/s and is plotted as the full line curve in Fig. 9, with the full-controlled three-phase curve dotted in for comparison.

Once more it is shown that the half-controlled rectifier has the advantage of requiring a finite inductance down to zero output; this is of real practical benefit since it means that if one can choose a choke of millihenry inductance numerically equal to 3·33 multiplied by the maximum value of load resistance in ohms—then one can be sure that operation will be above critical inductance for all values of α from full d.c. output voltage to zero output voltage. On the debit side, however, it will be noticed that the inductance required in the three-phase case is greater for the half-controlled case than for the full-controlled rectifier at all output settings down to about 12 per cent (where the crossover occurs). This may seem a surprising result in view of the complete inductance advantage obtained in the single-phase case; there is, however, a practical explanation which is not difficult to understand with the aid of Fig. 10. Fig. 10(a) shows the voltage waveform from a full-controlled rectifier when the firing angle is retarded sufficiently to give 50 per cent of full output ($\alpha = 60°$); it will be seen that the amplitude is less than the peak amplitude of the full output and that the fundamental frequency of the waveform is six times the supply frequency. In the case of the half-controlled rectifier of Fig. 10(b), however, for the same 50 per cent cut in output the amplitude is still equal to the full peak and the fundamental frequency is only three times the supply frequency. One would therefore expect to have to use a larger choke for the half-controlled rectifier than for the full-controlled in order to reduce the a.c. ripple to the same d.c. value.

Fig. 9. Comparison of critical inductance requirements for half and full-controlled three-phase bridge rectifiers



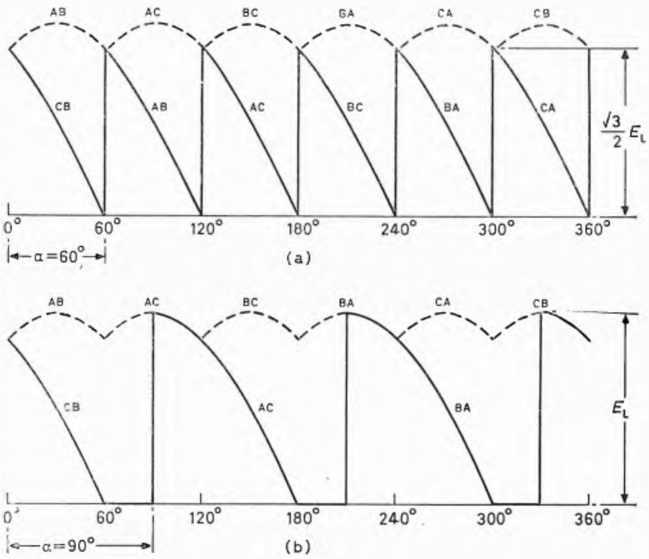


Fig. 10. Comparison of voltage waveforms at 50 per cent output
(a) Full-controlled three-phase bridge rectifier
(b) Half-controlled three-phase bridge rectifier

Comparing the critical inductance requirements for the single and three-phase half-controlled rectifiers it will be noted that, as might be expected, the single-phase circuit always requires more inductance than the three-phase; however, it is interesting that as α is increased the gap narrows from a ratio of about 34/1 at $\alpha = 0°$ to only 1·5/1 at $\alpha = 180°$. The ratio of 1·5 at 180° would appear to reflect the only significant difference between the waveforms as this angle is approached, i.e. their ripple frequencies are in the ratio of 150 to 100.

Peak Currents in the Thyristors

The foregoing equations include those for the instantaneous value of output current at any point in the cycle when the choke inductance is just critical for the particular value of load and firing angle. At any instant this current is flowing to the choke either from the supply via a diode-thyristor path in the bridge, or in a closed loop via the bypass diode. When the current is at its peak it will always be flowing through a thyristor path (rather than through the bypass diode) since, as already explained, the current peak occurs on the downward-going edge of the voltage sinusoid, i.e. before reaching the flat portion of the waveform—which corresponds to bypass diode conduction. Therefore it is only necessary to know the peak value of the equations for instantaneous choke current in order to determine the thyristor peak current. This involves finding the angle, ωt, at which the peak current occurs in each case (by setting the a.c. voltage function equal to the d.c. voltage) and then substituting this value of ωt in the appropriate equation for i. The results of such an analysis are given in Table 1.

The ratio i_p/I_{dc} is of course the ratio of the peak thyristor current to the d.c. load current. In a single-phase bridge therefore, the peak-to-mean thyristor current ratio can rise to a maximum of 4 when $L = L_c$ and, in a three-phase bridge, to a maximum of 6.

In practice it is unlikely that one would ever work exactly at critical inductance. One would generally design, or choose, a choke such that its inductance is always above the critical value for the anticipated extremes of load and firing angle. This means that under all normal working conditions there will be a reasonable safety margin above the critical inductance. It can be shown that if a is the value of i_p/I_{dc} which applies to the critical inductance case at any firing angle (see Table 1), then this ratio reduces to b for values of L greater than L_c where

$$b = 1 + \frac{a - 1}{\rho}$$

and ρ is the ratio L/L_c (greater than unity).

This enables one to calculate the peak-to-mean thyristor current ratio for practical values of inductance. For example, if 75A mean current is to be drawn from a three-phase bridge (containing 25A thyristors and diodes) when the output voltage is set to 25 per cent of maximum, then:

$$\alpha = 120° \text{ for 25 per cent output } \left(\frac{1 + \cos 120°}{2} = 0{\cdot}25\right)$$

$$i_p/I_{dc} = 1{\cdot}70 \text{ for } L = L_c \text{(from Table 1)}$$

If $L = 1{\cdot}5\ L_c$ at 120° is the practical case (the value of L_c being read from Fig. 9 at 120°, with the actual value of R at 120° being inserted), then

$$b = 1 + \frac{1{\cdot}7 - 1}{1{\cdot}5}$$
$$= 1{\cdot}47$$

Therefore the actual thyristor peak current for this condition is

$$1{\cdot}47 \times 75 = 110\text{A},$$

TABLE 1

	SINGLE-PHASE HALF-CONTROL		THREE-PHASE HALF-CONTROL	
α (degrees)	ωt for peak thyristor current (degrees)	i_p/I_{dc} when $L = L_c$	ωt for peak thyristor current (degrees)	i_p/I_{dc} when $L = L_c$
0	140·5	2	106·2	2
30	143·6	1·89	117·1	1·79
60	151·5	1·72	133·9	1·65
90	161·4	1·74	151·3	1·61
120	170·8	1·80	166·1	1·70
150	177·6	1·89	176·3	1·86
180	180	2	180	2

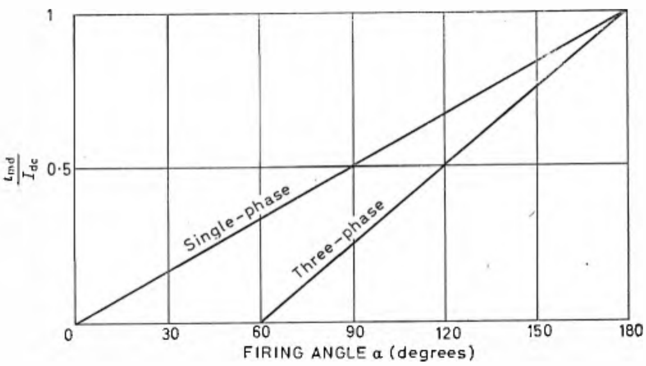


Fig. 11. Ratio $\frac{\text{mean current in bypass diode}}{\text{mean d.c. load current}}$ against α for infinite choke inductances (*Ratio is smaller for smaller inductances*)

which is within the peak repetitive current rating of, for example, CRS.25 thyristors (25A mean rating). (Note that, had the inductance just been at critical value, the peak repetitive current rating of 125A (for the CRS.25) would just have been exceeded: $1{\cdot}70 \times 75 = 127{\cdot}5$A.)

Peak and Mean Currents in the Bypass Diode

From Figs. 4 and 7 it is clear that the bypass diode begins to conduct for angles of α greater than 0° in the single-phase case, and for angles greater than 60° in the three-phase case. Above these values this diode takes more and more of the mean load current, and the thyristors and bridge diodes take less and less. Since the output current has always passed its peak by the time that the diode starts to conducts it follows that:

(a) The mean current in the bypass diode is greatest when the choke inductance is infinite.

(b) The peak current in the bypass diode occurs at 180° and 360° in the single-phase case and at 180°, 300° and 60° in the three-phase case, i.e. where the diode starts to conduct and at which point the output current is already decaying.

From (a) it may be deduced that the share of mean current taken by the bypass diode is proportional to the firing angle above 0° (single-phase) or 60° (three-phase). This is shown in Fig. 11, from which it is possible to calculate the highest possible mean current in the bypass diode as a fraction of the load current at any firing angle.

Referring to (b) one can calculate the peak current in the bypass diode by calculating the choke current at ωt = 180°, which is the point at which the current is switched to the bypass diode in both single- and three-phase cases. By combining equations for instantaneous choke current and d.c. current one obtains the relationship:

$$i_{pd} = I_{dc}\,(1 + (y/\rho))$$

where i_{pd} is the peak diode current (i.e., the choke current at ωt = 180°)

and I_{dc} is the mean d.c. load current, y is a function of α given below, and ρ is the inductance ratio L/L_c.

Single-phase Equation for y

For α = 0 to 35·5°:

$$y = \frac{\frac{1}{2} + ((\alpha/\pi) - \frac{1}{2})\cos\alpha - \dfrac{\sin\alpha}{\pi}}{\cos\phi + (1/\pi)(\alpha + \sin\alpha) + \dfrac{1 + \cos\alpha}{\pi}(\phi - \alpha - (\pi/2))}$$

$$\text{where } \phi = \sin^{-1}\left(\frac{1 + \cos\alpha}{\pi}\right)$$

for $\alpha = 35{\cdot}5°$ to $180°$:

$$y = \frac{\frac{1}{2} + ((\alpha/\pi) - \frac{1}{2})\cos\alpha - \dfrac{\sin\alpha}{\pi}}{\frac{1}{2}(\cos\alpha - 1) + (1/\pi)(\alpha + \sin\alpha)}$$

Three-phase Equation for y

For $\alpha = 0°$ to $60°$:

$y = -p$ (i.e., until α reaches 60°, the diode carries no current at all).

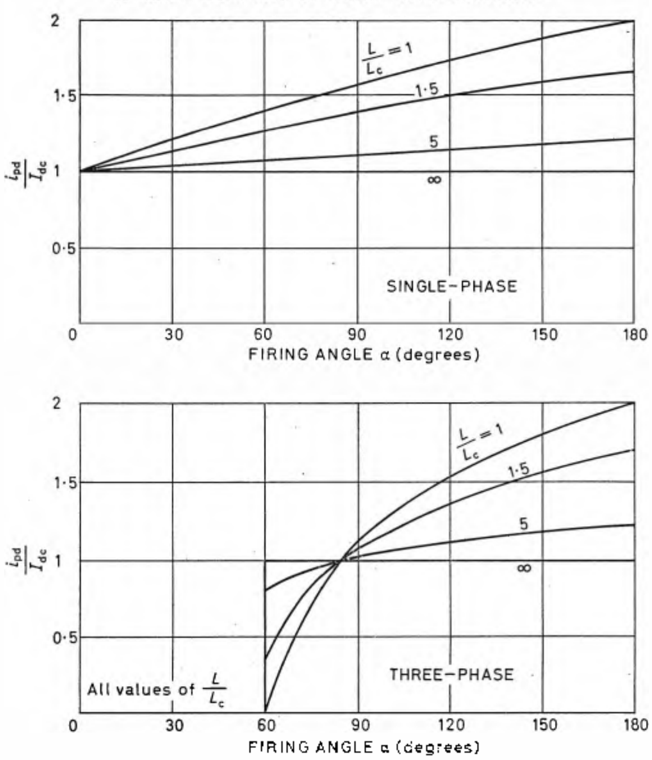
For $\alpha = 60°$ to $180°$ $$y = \frac{\frac{1}{2} + ((3\alpha/2\pi) - 1)\cos\alpha - (3/2\pi)\sin\alpha}{(3/2\pi)(\alpha + \sin\alpha) + \frac{1}{2}\cos\alpha - 1}$$

These equations for y have been evaluated and the resulting values of the ratio i_{pd}/I_{dc} have been plotted against α for both single- and three-phase cases in Fig. 12. It will be seen that in the worst case the bypass diode has a peak current equal to twice the d.c. load current (i.e. it must stand the same peaks as the thyristors).

Working Below Critical Inductance

Fig. 13 shows typical current and voltage waveforms from a half-controlled three-phase bridge when working at critical inductance (upper diagram) and at less than critical (lower diagram). In the critical case it will be noticed that the current just reaches zero at the firing point of the next thyristor (210° for example). If, however, the inductance is less than critical, the current will reach zero at some earlier point, as illustrated at *P* in the lower diagram. From *P* to the next firing point both the current and the rate of change of current are zero. From this it follows that the voltage across the choke must be zero over this period. Hence at point *P* the rectifier output

Fig. 12. Ratio $\dfrac{\text{peak current in bypass diode}}{\text{mean d.c. load current}}$ against α for various ratios of choke inductance to critical choke inductance



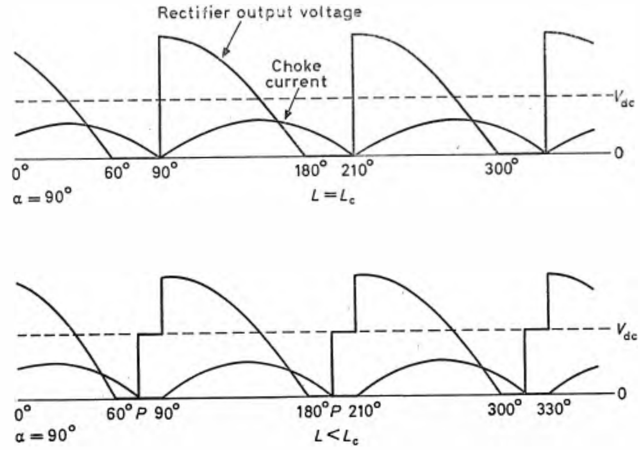


Fig. 13. The effect of using a choke of less than critical inductance on the voltage and current waveforms in a half-controlled three-phase bridge rectifier

voltage must shoot up to the level of the smoothed d.c. output voltage, thereby introducing a distortion of the rectifier output voltage waveform. This change of shape implies a rise in mean output voltage which may be interpreted in a number of ways. It can be regarded as equivalent to advancing the firing angle (by getting a higher output voltage—equivalent to firing earlier) or alternatively it can be regarded as bad regulation, i.e., the equivalent of feeding a capacitive instead of an inductive load. For large angles of α, where the thyristors only conduct for very short periods, the time between *P* and the next firing point can be a large part of the cycle and, since during this time no current flows, it is clear that large peak currents must flow during the short active periods of the thyristors in order to provide the required mean current output.

In many instances of course it will be found that it is quite uneconomical to provide critical inductance for much less than the maximum rated current output. This need not imply running the thyristors above their peak current ratings however, since the rise in peak-to-mean current ratio may be offset by the lower mean current taken by the load. However, under these conditions the natural regulation of the supply must suffer, and in an automatic voltage-regulated system the extra duty imposed on the control system should be taken into account.

Summary of Conclusions

(1) Full-controlled single- and three-phase bridge rectifiers have critical inductance values which tend to infinity as the rectifier output voltage is reduced towards zero; in contrast, half-controlled bridge rectifiers have values which remain finite right down to zero output voltage.

(2) In the single-phase comparison, see Fig. 5, the critical inductance is always lower for a half-controlled rectifier than for a full-controlled rectifier at any percentage-of-maximum output voltage setting. In the three-phase comparison, see Fig. 9, the full-controlled rectifier requires less critical inductance however for all output voltage settings greater than about 12 per cent of maximum.

(3) Comparing single- with three-phase half-controlled rectification, the critical inductance required for single-phase is considerably greater than that for three-phase at high output settings (34 times greater at maximum output voltage). At low output voltage settings, this difference reduces to a ratio of only 1·5 to 1 (compare Figs. 5 and 9).

(4) The peak current in each thyristor can reach twice the d.c. load current when the inductance is just critical; this means that in a single-phase bridge the peak-to-mean thyristor current ratio can reach 4, and in a three-phase bridge it can reach 6. These ratios are reduced for choke inductances greater than critical since the ripple current is reduced while the mean remains unaltered.

(5) The mean current in the bypass diode can equal the full d.c. load current if this current is to be drawn at very low output voltage settings. This means that the mean current duty of this diode may be twice that of each thyristor in a single-phase circuit, and three times in a three-phase circuit.

(6) The peak current in the bypass diode can equal twice the full d.c. load current if this current is to be drawn at very low output voltage settings. Its peak current duty can therefore equal that of each thyristor but cannot exceed it.

(7) Working below critical inductance must never be employed when running at the full rated mean current output of the thyristors or their peak current ratings may be exceeded; at light loads it may be permissible for economic reasons, i.e., to limit the choke size, to work below critical inductance provided that the peak current ratings of the thyristors are not exceeded. In this case the behaviour resembles that of a capacitive input filter and the voltage regulation becomes poor as soon as the load is reduced sufficiently to make the choke inductance less than critical. This regulation can, however, be taken up in automatic voltage control systems provided that proper attention is given to this point.

Acknowledgment

The author wishes to thank the management of Standard Telephones and Cables Ltd for permission to publish this article.

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Instrumentation Tape Recorders

A new family of instrumentation tape recorders, providing 2Mc/s bandwidth and more than 100 times the time-base accuracy of recorders currently in use, is being manufactured by Ampex International.

The new range consists of the model FR-1600, a full-size recorder for exacting ground station and laboratory use; portable model PR-1600 for field, van or shipboard application; and airborne model AR-1600 for use in aircraft at altitudes up to 70 000ft.

Two major design innovations contribute to these advances in accuracy. A dual-response capstan servo provides maximum tape stability during record and playback. A large-capstan tape drive design increases short wavelength response and tape support to minimize inter-channel time displacement error and flutter. Time-base error on the FR-1600 is as little as $\pm 0{\cdot}5\mu$sec at 120in/sec, compared with 250μsec on previous instrumentation recorders.

The FR-1600 provides 33 per cent more uninterrupted recording and playback time than previous instrumentation recorders by permitting use of 16in tape reels. Previous recorders of this type accommodate up to 14in reels. Playing time at 120in/sec is 20min, compared with 15min on 14in reels.

The 1600 range is the first integrated family of laboratory portable and airborne recorders to be developed simultaneously. This assures that tape interchangeability and signal integrity may be maintained when acquiring and reducing data under various conditions and provides new versatility for a wide range of recording applications.

The new capstan design, using only one rotating element, eliminates pinch rollers and rotating guides on the FR-1600, offering more stable tape motion and thus greater data reliability. Vacuum buffer chambers also give the FR-1600 gentle tape-handling ability, further reducing time-base error.

A new, compact digital tape transport which can record more than twice the amount of data on the same amount of tape used with previous recorders of its type is also being marketed by Ampex International. This is the model ATM-13 which is designed for airborne, shipboard and land-mobile use in such applications as reconnaissance and geophysical work. It is the first recorder of its type to generate data which is computer-compatible, requiring no intermediate processing. Blocks of data recorded by the ATM-13 are spaced only ½in apart on the tape, compared with distances ranging from two inches to several feet on comparable recorders currently in use.

The ATM-13's short inter-block gap meets IBM and other computer requirements for tape inputs to processing systems. The short gap, made possible by the transport's start-stop time of six thousandths of a second, offers users substantial savings in tape and processing time.

A servo-driven push-pull capstan drive, based on the single-capstan drive design pioneered by Ampex, results in the fast start-stop tape action. By minimizing tape strain, the design also increases data reliability.

Models include airborne, land-mobile and shipboard designs, seven-track or nine-track, either as a basic transport or with full data electronics.

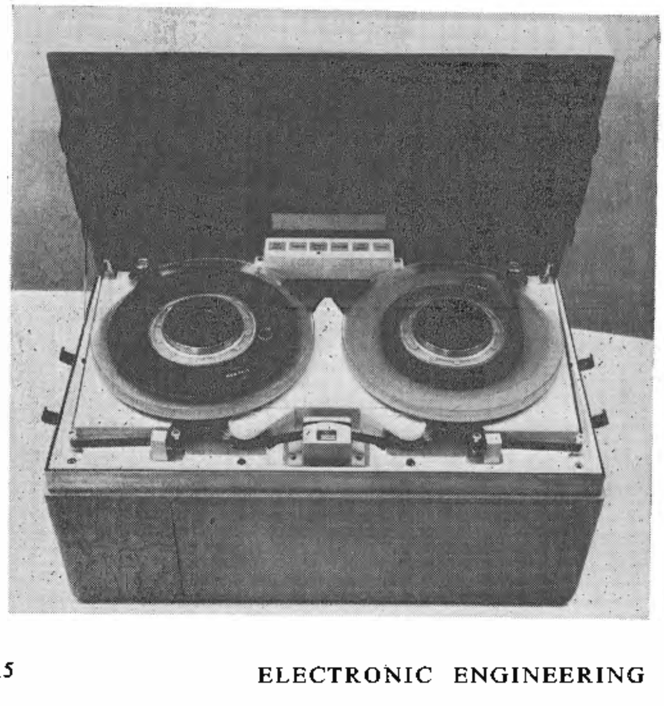
The specially designed transport, able to withstand shocks of up to 15*g*, operates reliably at altitudes of up to 70 000ft. Maximum tape speed is 75in/sec in read/write, start/stop mode, and more than 100in/sec in continuous read and/or write (gapless) mode. In either mode, the ATM-13 can pack up to 800 characters on an inch of magnetic tape.

Weight of the transport and separate power supply/control totals 150lb. The transport measures approximately two feet long by one foot wide by one foot high. The power supply is approximately three-quarters that size.

Another new model is the FR-1800, an instrumentation recorder combining a high performance transport with a fast response servo, zero loop capstan and pneumatic chute guides. The 1·5Mc/s FR-1800 is said to offer 60 per cent fewer parts than any transport with comparable performance.

No mechanical adjustments and less than ten 'set and forget' electrical adjustments are required on the FR-1800 transport. Tape threading is simple and fool-proof—an added benefit of zero loop design. Built-in optional shuttle capability simplifies searching for data, and makes it possible to repeat data between two preset points.

The ATM-13 tape transport for airborne, land-mobile and shipboard use



A Transistorized Pulsed Drive for a Stepping Motor

By W. J. Jirafe*, B.E., M.Tech., Ph.D.

The article describes the development of transistorized circuit for driving a standard four-lead stepping motor either clockwise or counterclockwise by pulses to its input. The motor advances a step per pulse and the sequence is controllable by means of a switch. The switching sequence is generated by two flip-flops, the outputs of which, on amplification, drive the two phases of the motor. They are triggered by a common flip-flop actuated by input pulses through a sequence logic. The circuit is self-guarding against mains and other disturbances. An alternate scheme employing a reversible binary counter is suggested and the favourable features of the former in comparison with the latter are illustrated. Arrangement for a motor with bifilar windings is indicated. A brief study of the mode of operation of a stepping motor and its application as a position servo is also given.

(Voir page 345 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 352)

STEPPING motors are based on the simple 'solenoid' electromagnet in which a ferromagnetic armature moves in the field of a coil carrying a current. It can be shown[1] that the armature tends to move in such a direction that the flux linked with the exciting coil is increased.

Fig. 1(a) illustrates a motor with a four-coil stator and a unipole five-tooth rotor. Fig. 1(b) gives its drive circuit, the switching sequence of which is shown in Table 1.

TABLE 1
Switching Sequence

Clockwise Rotation			Counterclockwise Rotation		
Rotor Position	Phase A (S_1)	Phase B (S_2)	Rotor Position	Phase A (S_1)	Phase B (S_2)
1	+	+	1	+	+
2	−	+	4	+	−
3	−	−	3	−	−
4	+	−	2	−	+
1	+	+	1	+	+

The rotor positions shown in Fig. 1(a) correspond to the positions indicated in Table 1. Each time the polarity is changed, the rotor assumes the new position of maximum permeance. It is seen that the rotor advances by 18°, i.e., by the vernier difference between the stator and the rotor pitches ($\theta_u = 360(1/4 - 1/5) = 18°$) per step, and the motor makes 20 steps per revolution. Had there been no difference in the pitches, the number of steps per revolution would have been just 4.

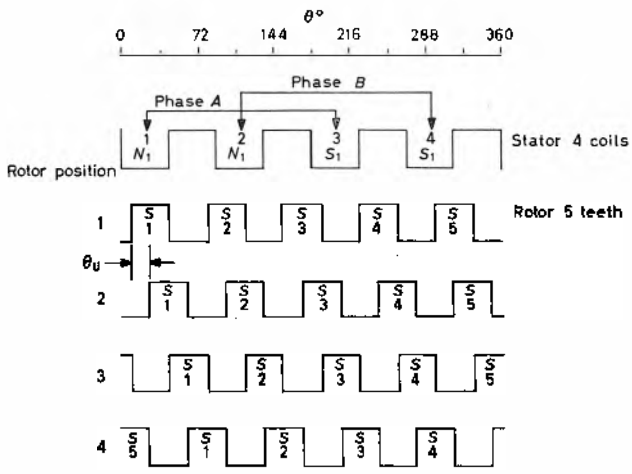
The switching sequence is analogous to the operation of a two-phase motor in which the two windings A and B are excited in phase quadrature as shown in Fig. 2. The four positions corresponding to the switching sequence above are indicated by dots. Each time, in this case, the magnetic field advances by 90°.

These motors have also been referred to as 'restricted motion motors' and also as 'incremental motors'. Their use as a stepping or incremental positioning device offers control system designers, machine tool and process control engineers unlimited application possibilities.

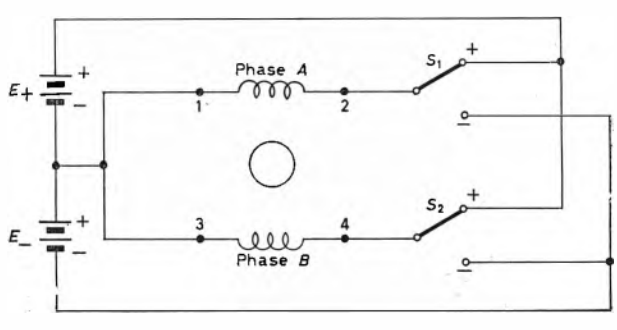
Stepping Versus Stepless Control

Fig. 3 compares a position servo with its incremental counterpart. A stepping motor may be viewed as the simplest form of incremental servo[2], i.e., for each input signal its output shaft assumes a new increment of angular position. Physically and analytically this effect resembles the feedback of a conventional position servo. It could be argued that a stepping motor is not a true position servo because it does not measure the position of its output shaft, feed back a signal proportional to this measure-

Fig. 1. A stepping Motor
(a) Rotor advances in steps
(b) Drive circuit using switches



(a)



(b)

* *Indian Institute of Technology, Bombay, India.*

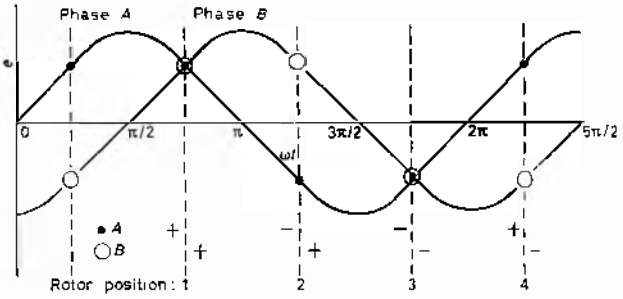


Fig. 2. Analogy with a two-phase motor

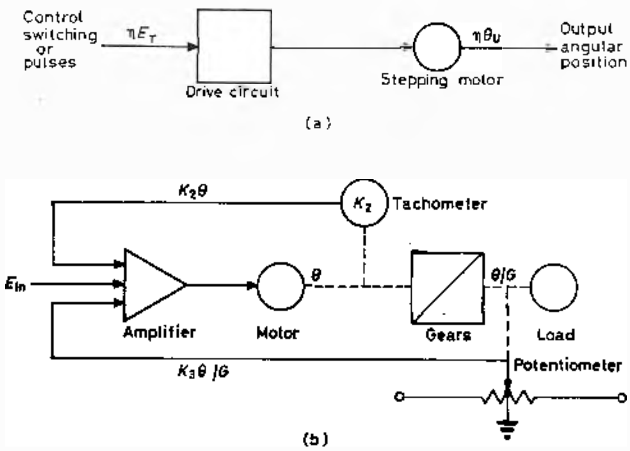


Fig. 3. Two types of position servos
(a) Incremental position servo
(b) Conventional position servo

ment, compare it to the input signal, and then apply the difference to torque the output shaft to its correct position. But it must be noted that if command information were available in small sign-sensitive 'bits' to which the stepping motor could respond with high incremental accuracy and stability, the motor would then perform the basic function of the position servo within the limits of its frequency response and incremental accuracy. Salient features of this scheme are:

(a) A suitable stepper may replace amplifier, servomotor, tachometer, gear train and transducer and still offer all the desirable features of a feedback system without hunting or instability.

(b) Power consumption for intermittent operations can be reduced to zero during quiescent periods, a feature useful in the design of missile and space vehicle controls.

(c) Use of an essentially digital system is far more compatible with advancing digital data techniques, both digital and analogue outputs are available, though, as well as the discrete step mechanical output.

Slo-Syn Motor

The stepping motor for which a transistorized drive has been developed is a 'Slo-Syn' motor type SS50-1011 of Superior Electric Company, U.S.A. It has 8 coils with teeth cut in the pole pieces to give, in effect, 48 teeth. The rotor has 50 teeth and is of unipolar type. It consists of two sections, one north and the other south phased by ½ pitch so that the same windings could be used for both the sections. Alnico permanent magnet is used for magnetizing the rotor.

There will be 6 teeth per coil of the stator and all 6 teeth will bear the same polarity for a particular step. The rotor, therefore, will advance 6 vernier differences per step ($\theta_u = 6 \times 360\,(1/48 - 1/50) = 1{\cdot}8°$) so as to attain the position of maximum permeance as explained earlier. The number of steps per revolution will be 200.

SPECIFICATIONS

Four-lead 'Slo-Syn' Stepping Motor
Type SS50-1011

D.C. volts	= 10·0
Ampere/winding	= 0·75
No. of steps/rev.	= 200
Accuracy per step (1·8°)	= ±0·09° non-accumulative
Starting torque (oz-in)	= 50
Maximum speed (steps/sec)	= 300
Weight (lb)	= 3

Logical Design

The pulsed drive, instead of switches shown in Fig. 1(b), uses suitable logic circuits enabling it to force currents in accordance with the sequence of Table 1 in phases A and B simply on the application of pulses to its input. The motor advances by a unit step per pulse and the sequence of rotation is controllable by means of a switch used for feeding binary levels pertaining to two sequences: clock-

Fig. 4. Arrangement of the pulsed drive

F_t = *triggering flip-flop, output* T, T'

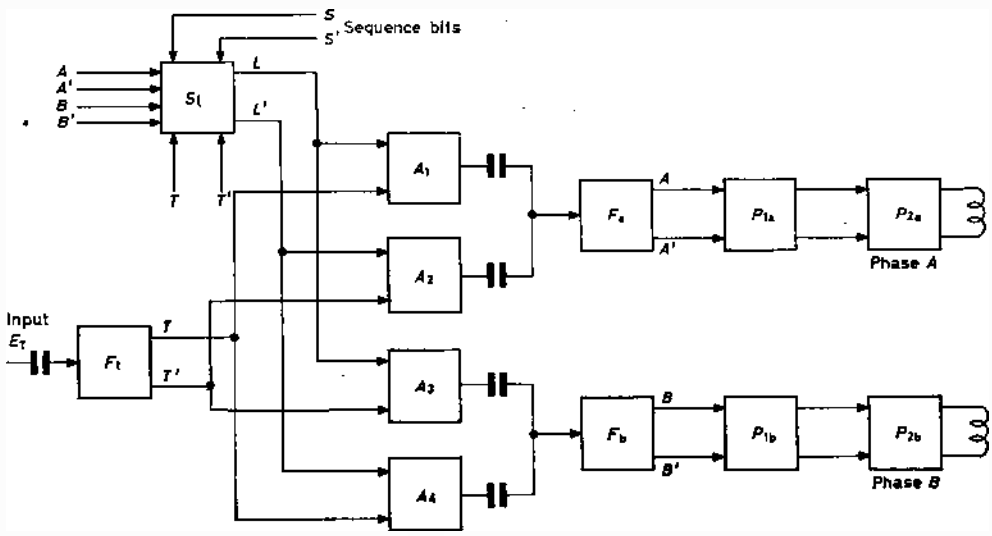
F_a, F_b = *switching flip-flops, outputs* A, A' *and* B, B'

S_1 = *sequence logic, output* L, L'

A_1-A_4 = *AND gates*

P_{1a}, P_{1b} = *preamplifiers*

P_{2a}, P_{2b} = *power amplifiers*



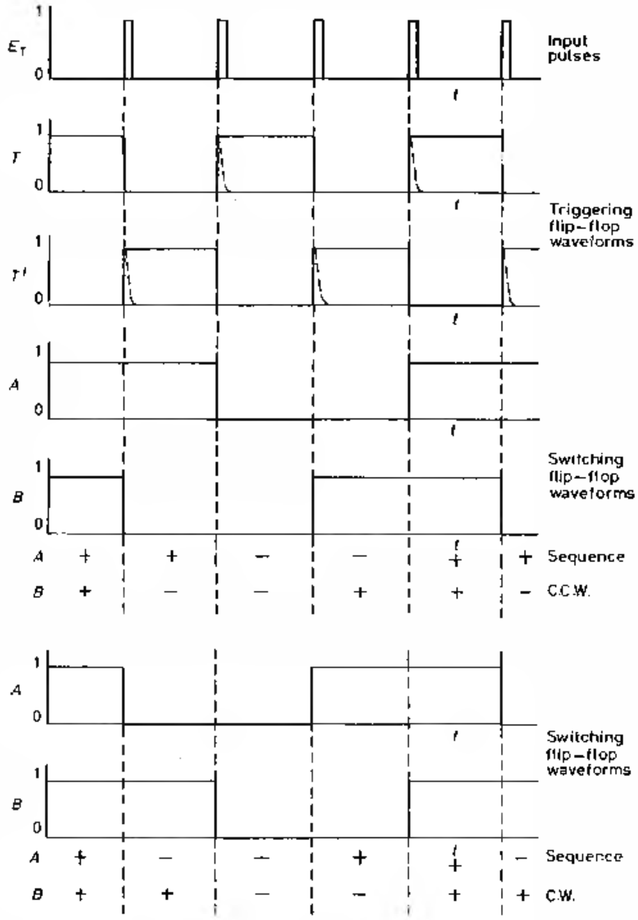


Fig. 5. Logical waveforms

wise and counter-clockwise. For controlling the sequence in accordance with a change in signal voltage level, it may be replaced by a suitable Schmitt trigger.

Outline of Scheme

A bistable multivibrator shown as a triggering flip-flop, F_t in Fig. 4 forms an input of this drive circuit. Input pulses are applied to this flip-flop, and depending upon initial states and the sequence desired, its outputs are used to trigger either of the switching flip-flops F_a and F_b. Sequence logic S_1 and AND gates A_1 etc., carry out discrimination. Pre-amplifiers P_{1a} and P_{1b} and power amplifiers P_{2a} and P_{2b} suitably amplify their outputs and feed the phases A and B. Logical waveforms of the outputs of these flip-flops are shown in Fig. 5. It is assumed that the phase current is positive when A or B is 1 and negative when it is 0. Also a flip-flop makes a transition only for a positive-going trigger.

TABLE 2
Truth Table

S	1	1	1	1	0	0	0	0
X	1	1	0	0	1	1	0	0
A	1	0	1	0	1	0	1	0
B	0	1	1	0	1	0	0	1
L	1	1	1	1	1	1	1	1

It is obvious that when a pulse is applied, only one of the two switching flip-flops will change its state. This is what is required by the switching sequence given in Table 1. With the initial states of F_t, F_a and F_b shown in Fig. 5 ($T = 1$, $A = 1$ and $B = 1$), it is seen that when T triggers F_a and T', F_b, the sequence is counter-clockwise whereas when T triggers F_b and T', F_a, i.e., vice-versa, the sequence is clockwise. It should be understood that the choice of triggering for a particular sequence will depend upon the combination of initial states.

Sequence Logic

With T triggering F_a and T', F_b, analysis of waveforms reveals that an even number of unity initial states gives clockwise sequence whereas an odd number of unity initial states gives counter-clockwise sequence. Obviously, for all the combinations corresponding to this observation, the output of the sequence logic should be $L = 1$ (and $L' = 0$). In all other remaining cases, it must come out as $L = 0$ so that T triggers F_b and T', F_a. Taking these facts into account, a truth table can be formulated (Table 2). Sequence bit $S = 1$ for clockwise and $S = 0$ for counter-clockwise sequence.

An Exclusive OR function[3] (symbol $a\,(+)\,b$) provides a simple as well as an economical solution to this problem. Taking:

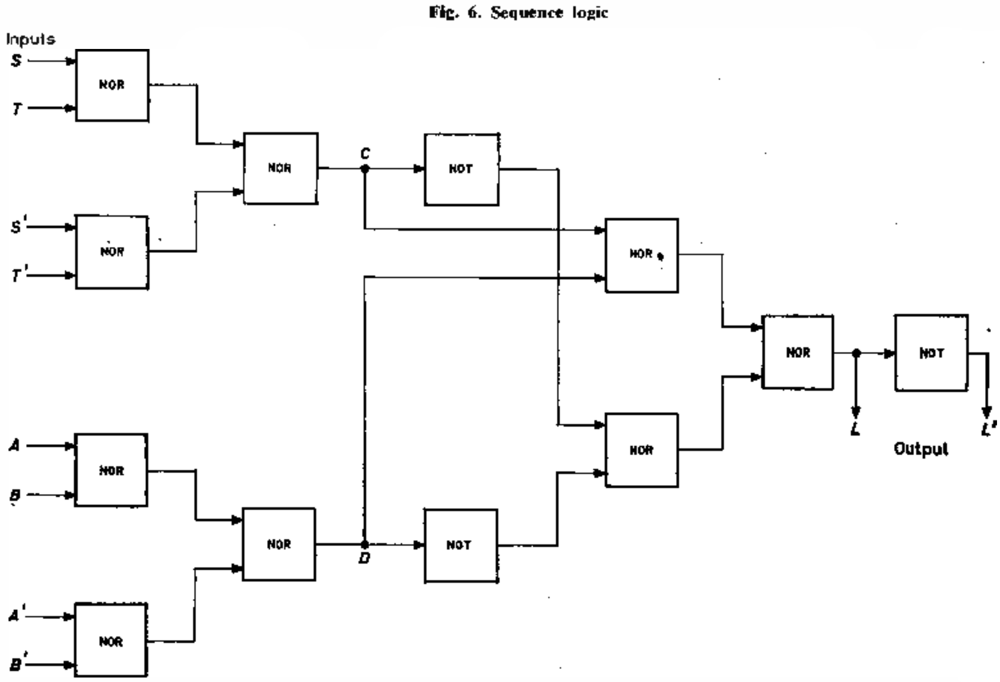
$$C = S\,(+)\,T,$$

and

$$D = A\,(+)\,B$$

$$L = C\,(+)\,D \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Fig. 6. Sequence logic



Boolean expressions used for realizing these functions are:

$$C = (S + T)(S' + T'),$$
$$D = (A + B)(A' + B')$$

and

$$L = (C + D)(C' + D') \quad \ldots\ldots\ldots\ldots \quad (2)$$

T.D.L. (Transistor-Diode Logic) NOR circuits[4] have been used for their synthesis. Reduction to the desired form can be obtained as follows:

$$C' = [(S + T)(S' + T')]'$$

By De Morgan's theorem

$$C' = (S + T)' + (S' + T')'$$

Now:

$$C = C'',$$

or

$$C = [(S + T)' + (S' + T')']' \quad \ldots\ldots \quad (3)$$

Each Exclusive OR function requires three NOR circuits and will be followed by a NOT circuit for inversion. The arrangement of the complete sequence logic is given in Fig. 6.

Circuit Design

Amplifier Circuits

The power amplifier of the phase A channel of Fig. 7 replaces the two-way switch S_1 of Fig. 1(b) by power transistors VT_{31} and VT_{32}. Depending upon the state of the switching flip-flop F_a, one of them is saturated and the other is cut-off. When VT_{31} is cut off and VT_{32} is saturated, say, the phase current is positive, with VT_{31} saturated and VT_{32} cut off, it will be negative. These phase windings are highly inductive and generate high voltage transients which may damage the transistors. Power transistors with excess voltage ratings (2N174, collector-to-base voltage $\simeq -80$V max.) have, therefore, been used.

Average flip-flop output voltage levels are -6V and 0V. Save for the AND gates, the logic circuits have been designed with -6V for 1 and 0V for 0. When $A = 1$, the pre-amplifier has VT_{21} cut off and VT_{22} saturated. Correspondingly, VT_{31} is cut off and VT_{32} saturated and the phase current is positive. The base-to-emitter voltage of the cut off transistor is $+5$V and the base current of the saturated transistor is -40mA. When $A = 0$, the conditions will be reversed and the phase current will be negative.

Power supplies for these circuits are also shown in Fig. 7. All are centre-tapped bridge full-wave rectifier circuits. Levels E_- and E_+ use choke input filters while the rest of the levels use RC filters and are Zener-stabilized. Stepping motors with bifilar windings (each phase has two halves with one half wound in opposite direction; there are 5 leads coming out) would require a less number of supply levels as the reversal of phase current is obtainable merely by introducing the two halves as collector loads of a flip-flop. But since such motors have twice as many winding turns as the standard type, for the same size of motors, the low speed torque is reduced.

Logic Circuits

Input and logic circuits have been shown in Fig. 8. Capacitors across the collector loads of L and L' generating circuits are necessary for by-passing spurious pulses arising out of difference in delays introduced by the transistors in the circuit.

AND gates are analogous to the OR gates of the NOR circuits and, therefore, have input voltage levels interchanged (0V for 1 and -6V for 0). Since their output

Fig. 7. Phase 'A' channel

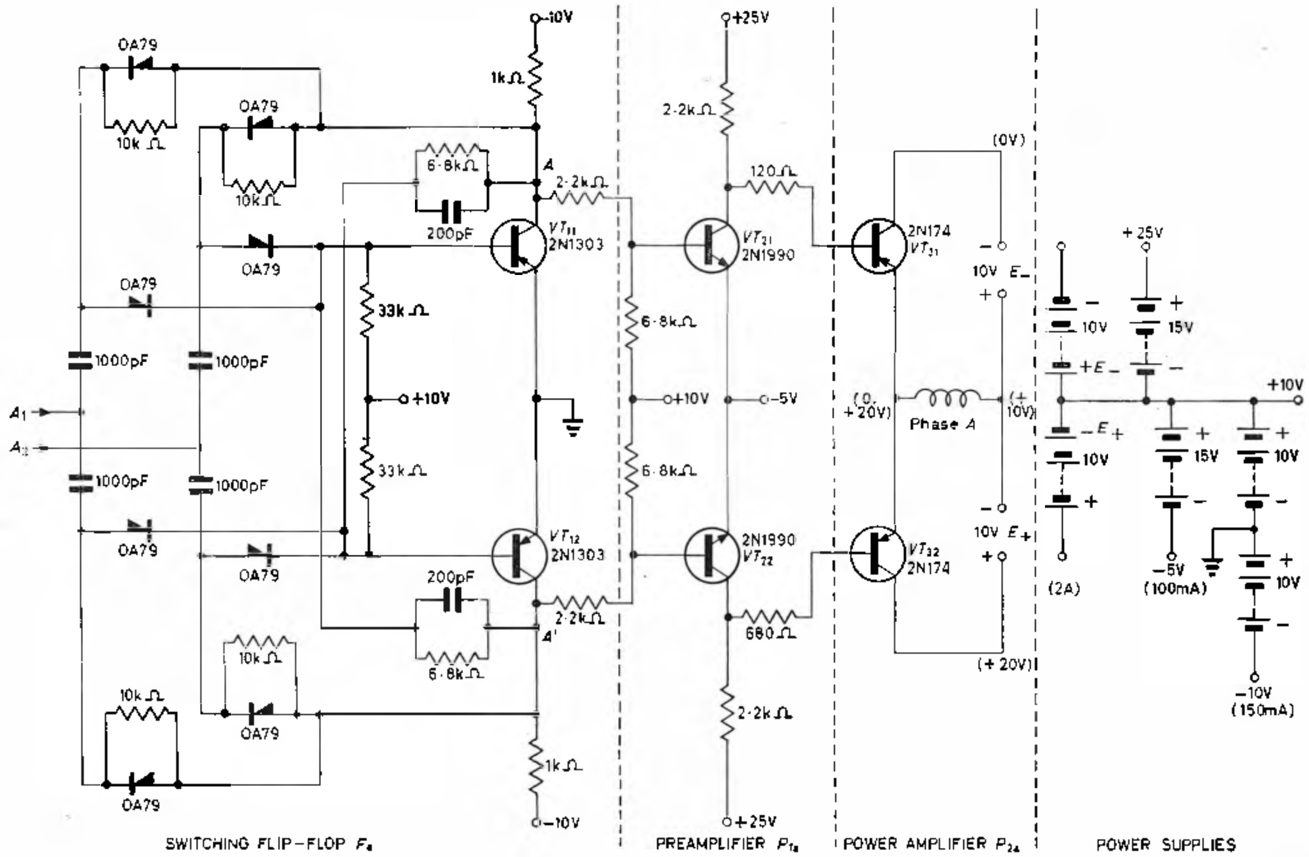


TABLE 3
Reversible Binary Counter

Input Pulses/ Rotor Position		Counter Output		Desired Output for Switching Sequence	
$CW\downarrow$	$CCW\uparrow$	A	B	Phase A (A)	Phase B (B_0)
1	1	0	1	0	0
2	4	1	0	1	0
3	3	1	1	1	1
4	2	0	0	0	1
1	1	0	1	0	0

follows the most negative input voltage, only that AND gate on which the sequence logic output is 0V will transfer the triggering flip-flop output and trigger the switching flip-flop. In Fig. 8, when L is 0V, T will trigger F_a and T' will be blocked as L' will be -10V. When L reverses, T' will trigger F_a. Referring to the earlier discussion, it may be noted that owing to the reversal of signal levels for AND gates, merely the role of T and T' is interchanged.

Performance

The drive was actuated by a square wave and the waveforms were observed on an oscilloscope. The performance was uniformly satisfactory from 0·1c/s to 1kc/s for both of the sequences. The input signal amplitude required was about 3 to 5V peak-to-peak. The stepping motor was also found to step satisfactorily without losing a single step for either of the sequences from 0·1c/s, but up to only 300c/s on the higher side. Between 300c/s and 350c/s, the performance was uncertain and beyond 350c/s it was certain in losing steps. In the useful range, the changeover from one sequence to the other was sudden and the performance was unaffected by mains disturbances. No specific test to study the effect of temperature variation on performance was carried out, but the system was found to work uniformly for hours together for any part of a day.

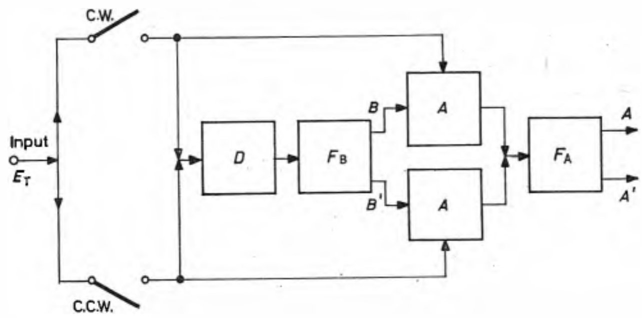


Fig. 9. Reversible binary counter

An Alternate Scheme

A reversible binary counter[5] shown in Fig. 9 could also be used for generating sequential signals for the drive. The mode of generation is given in Table 3.

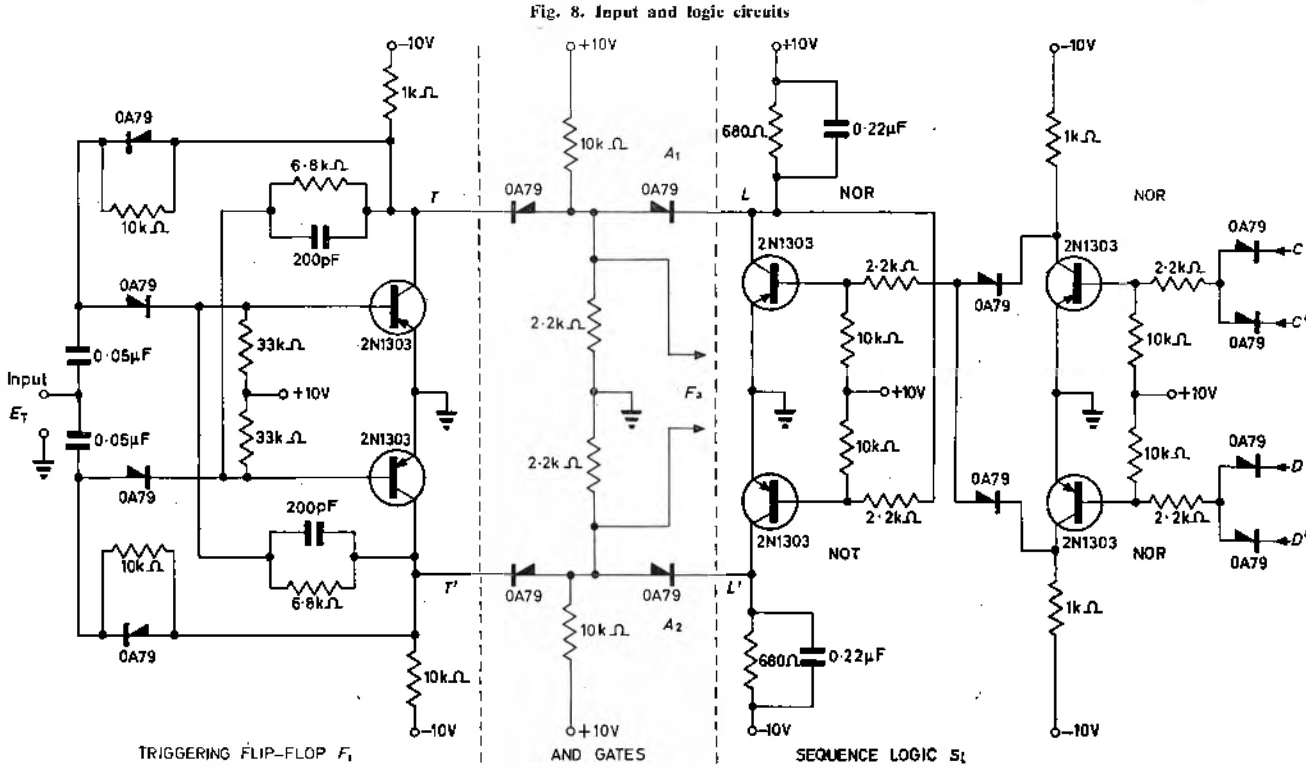
Assuming phase current as positive for 1 and negative for 0 as before, it is seen that while A can directly be used for phase A, for phase B, B has to be modified as B_0 which may be obtained from the expression:

$$B_0 = AB + A'B' \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

A and A' and B_0 and B' should now be fed to the respective amplifier circuits to obtain the desired phase currents.

The above scheme, although it appears simple and straightforward, has certain drawbacks. It can be seen from Table 3 that at alternate positions, both of its flip-flops change their states simultaneously and that the desired output B' is obtained through logic circuits. Owing to the deliberate delay at the input to the flip-flop F_B and also to that added by the transistors in the logic circuits,

Fig. 8. Input and logic circuits



this gives rise to intermittent pulses which are likely to cause spurious steps. The former scheme, on the other hand, is thoroughly free from such disturbances as at any time, only one of its switching flip-flops changes its states.

Further, since its sequence logic takes into account the initial states of all the flip-flops in the circuit, unlike the reversible binary counter, the scheme is self-guarding against mains and other disturbances. Over and above these favourable features, the two schemes require more or less the same number of components.

Acknowledgments

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A Telemetering Transducer System

The 'Tele-Meter', manufactured by I.E.C. (Electronics) Ltd, Walton-on-Thames, Surrey, is a commercial version of an instrument designed and developed by the Motor Industry Research Association. It is intended for use where signals from strain gauges or other transducers have to be obtained from moving parts and where the attachment of wires or the use of slip-rings or other rotating contacts is difficult or inadvisable.

Originally designed for the measurement of torque in vehicle propeller shafts it consists, basically, of a small frequency modulated transmitter and aerial which is attached to the shaft, a receiving aerial which is placed a few inches away from the shaft and the receiving, discriminating and recording unit which may be placed in a remote position.

The parts attached to the shaft are small and light; the frequency response extends from d.c. to at least 100c/s in its present form, but the upper limit can be substantially increased. The sensitivity is such that full scale deflexion is obtained for 0·1 per cent strain when wire-resistance gauges are employed, and the apparatus can be used over the temperature range -5 to $+35°$C.

The choice of the nominal centre frequency was a compromise; the signal transmission requirements suggested a relatively high centre-frequency, whereas the first modulator circuit that was tried required the strain gauge supply to be at the carrier frequency, which would have led to capacitive balance problems if the frequency had been too high. As an existing discriminator was available for a centre frequency of 6kc/s (a standard frequency in f.m. tape recorder practice), this frequency was chosen as the nominal operating frequency of the modulator.

Three types of circuit were considered:

(1) A d.c. amplifier feeding into a frequency modulator.
(2) An oscillator providing the strain gauge supply, and a.c. amplifier and phase sensitive discriminator feeding into a frequency modulator.
(3) An oscillator in which the strain gauge bridge formed part of a secondary feedback loop.

The first and third type of circuit appeared to be the most suitable for development, the second type being discarded as too complex and unlikely to satisfy the size and weight requirements. Initially experiments were carried out with the third type of circuit, which offered the possibility of high gain without the use of multi-stage d.c. amplifiers with their associated drift problems.

The type (3) circuit, which has been used elsewhere for strain measurement, but not necessarily in miniature form, is attractive because the gain of the amplifier within the feedback loop is not critical as long as it is high enough to maintain oscillations. Although a working model employing the principle was constructed, it was found necessary to introduce gain in the auxiliary feedback loop, in which the stability of the amplifier was critical. The final circuit was relatively complex and, of necessity, incorporated relatively heavy and bulky iron-cored components and variable capacitors. It became clear that this type of circuit would not meet the size and weight restrictions.

Attention was therefore turned to obtaining optimum performance from a type (1) circuit. It was found that a high sensitivity modulator could be made, which preceded by a single stage of d.c. amplification provided adequate sensitivity to measure torsional strain of the order of 0·1 per cent, in conjunction with wire resistance strain gauges; higher sensitivities could be obtained when required, by employing semiconductor strain gauges. It was accepted that this approach represented the best compromise for this application, and the final circuit is based on the type (1) circuit. Several versions were constructed and used under actual operating conditions before deciding upon the final design.

Initially a mains operated discriminator was used in conjunction with the modulator, but as the development of the modulator progressed and road tests became necessary, the need for a small battery operated discriminator became apparent. Accordingly, a discriminator was made using conventional circuits which fed into a galvanometer recorder. Since in some applications, the cable from the receiving 'aerial' to the discriminator may need to be fairly long, a pre-amplifier with a low output impedance was built into the receiving 'aerial'.

To protect the modulator from damage and to support the components under high centrifugal accelerations that exist when used on a high speed shaft, it is potted in epoxy resin and electrical connexions are made by means of pins which protrude from the casting. The battery box is constructed from Tufnol to match the modulator in size and weight, and the two units can be mounted on a shaft by means of a split ring which also acts as the transmitting aerial.

The 'Tele-Meter'

Thermistor Networks for D.C. Stabilization of Transistors

By H. Singh*, M.Sc., M.Tech., M.I.E.E.

An analytical approach to the design of thermistor networks for the stabilization of d.c. operating point of transistors has been developed. It is based on the principle of thermistor compensation techniques which enable the stabilization of collector current by adjusting the base-to-emitter voltage of the transistor at three pre-selected temperatures. The method described gives considerable improvement in the d.c. collector current stabilization, for the temperature range of 20 to 50°C.

This method serves as an alternative to graphical methods for shaping the characteristics of thermistors.

(Voir page 245 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 252)

THE effect of temperature on the shift of d.c. operating point of transistors was observed in the early stages of development of transistor circuits. This shift is primarily caused by the changes in transistor parameters with ambient temperature. With increase in ambient temperature, collector current of a transistor increases, resulting in a reduction of signal handling capacity, and if the temperature increases beyond a certain limit, it may lead to a thermal breakdown of the transistor. Attempts have been made from the early stages of the development of transistor circuits to stabilize the collector current against variation in ambient temperature[1]. Linear as well as non-linear circuit elements have been tried for the d.c. stabilization of collector current. The most common method using linear elements is by means of a potential divider and emitter resistor[2]. Although this method has proved to be of some use for general purpose work, the disadvantages are that it leads to a lower input impedance and a higher current drain from the supply. The other common method for stabilization is the use of a thermistor for compensating the effects of temperature variation[3].

Though thermistors have either negative or positive temperature coefficients, common use is limited to those having negative temperature coefficients and these are the types that have been used here. The stabilization of collector current by using thermistors removes both the disadvantages of the potential divider and emitter resistor method and also gives a higher degree of stability.

The d.c. operating point for a common-emitter configuration of transistor is defined by means of collector current and collector-to-emitter voltage on the output characteristic curves of the transistor. There are three main causes for the variation of the d.c. operating point with ambient temperature, namely, the effects of changes of reverse saturation current, I_{CBO}; base-to-emitter voltage $V_{B'E}$ (B' denotes the intrinsic base); and d.c. collector current amplification factor, h_{FE}.

Variation of D.C. Parameters with Temperature

The variations of d.c. parameters with temperature have been reported by several workers[4-11].

Neglecting the temperature variation of the energy gap, the rate of doubling of the reverse saturation current I_{CBO}, with temperature is given by[12]:

$$\Delta T = \frac{T}{5/2 + (qV_g/kT)} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

where ΔT = rise in junction temperature required for doubling the reverse saturation current

T = absolute value of the temperature (°K)

q = electronic charge

k = Boltzmann's constant

V_g = energy gap of the semiconductor

The rate of change of reverse saturation current, I_{CEO}, is higher than that of I_{CBO} due to the fact that at a lower level of current injection, d.c. collector current amplification factor, h_{FE}, also increases with temperature.

The rate of change of base-to-emitter voltage, $V_{B'E}$,

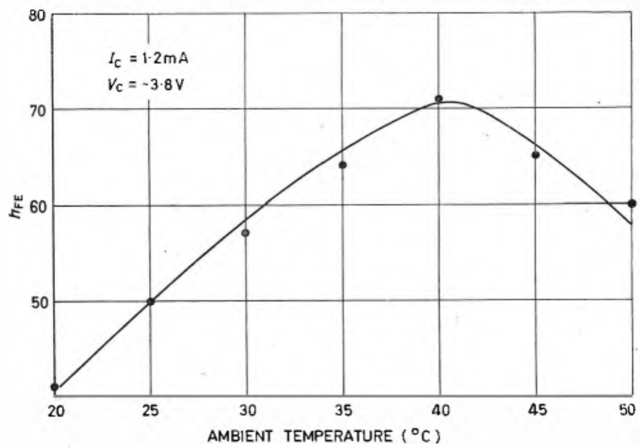

Fig. 1. Variation of h_{FE} of Mullard type OC71 transistor with temperature

with temperature for constant emitter current is given by[13]:

$$\left.\frac{\partial V_{B'E}}{\partial T}\right|_{I_E} = -\frac{V_g - V_{B'E}}{T} - (5k/2q) \quad \ldots\ldots (2)$$

There is no such simple relation for the variation of d.c. collector current amplification factor, h_{FE}, with temperature. But its variation with temperature can be experimentally measured. The nature of the variation of h_{FE} with temperature is almost the same for all types of transistors. Fig. 1 shows the variation of d.c. collector current amplification factor over the temperature range of 20 to 50°C for the Mullard type OC71 transistor in common-emitter configuration.

The design of the thermistor biasing network using graphical procedure[14] is sometimes inconvenient due to the fact that the values of the components of the network depend upon the position of the meeting point of two straight lines which may lie far away from the graphical space, because, to start with, one does not know before hand the position of the meeting point to choose the correct size of the graph, thus making it difficult to note accurately the meeting point.

The method described below is purely analytical. Here the biasing network is designed so that it tries to adjust the base potential of the transistor at three pre-selected values of temperatures in such a way that the increase in

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d.c. collector current with temperature is exactly zero at the selected temperature values.

Analysis of the Biasing Network

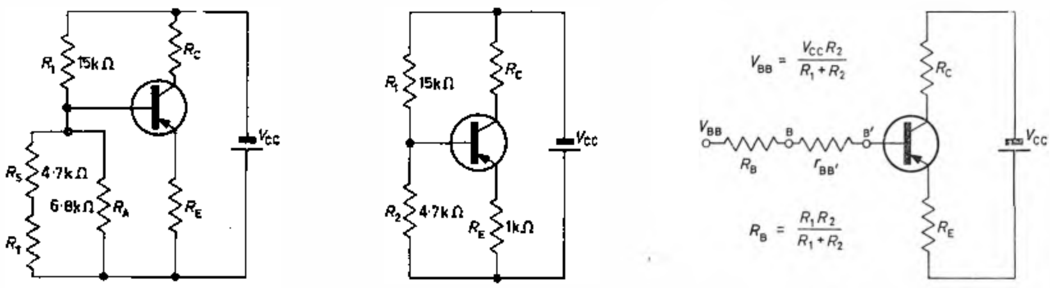
The principle of this method is explained by the help of Fig. 2. The thermistor, denoted by R_T, is used in the lower arm of the potential divider network. The series resistance, R_A, and shunt resistance, R_S, of the lower arm help to shape the variation of base potential to exactly nullify the increase in collector d.c. at three selected temperatures.

For the purpose of analysis, all other resistances except R_T have been assumed to have zero temperature coefficient. The d.c. collector current, I_C, is related to the other circuit elements and voltages and currents of the transistor by the following relation.

$$I_C\,[R_E\,(1+h_{FE})+R_B+r_{BB'}] = I_{CEO}\,(R_E+R_B+r_{BB'}) + h_{FE}\,(V_{BB} - V_{B'E}) \quad \ldots\ldots\ldots\ldots (3)$$

where R_E = external emitter resistance

R_B = Thevenin's equivalent series resistor of the biasing network



(a) (b) (c)

Fig. 2. Biasing circuit for a transistor

(a) Actual circuit

(b) Equivalent biasing circuit

(c) Thevenin's equivalent circuit

$r_{BB'}$ = base spreading resistance of the transistor

V_{BB} = Thevenin's equivalent supply voltage for the base of the transistor.

Partially differentiating the collector current, I_C, of equation (3) with respect to reverse saturation current, I_{CEO}, base-to-emitter voltage, $V_{B'E}$, d.c. collector current amplification factor, h_{FE}, and Thevenin's equivalent resistor, R_B, respectively, and adding them, one obtains the total change in the collector current, when all the above quantities are varying simultaneously, by the following relation.

$$\Delta I_C = S\left[\Delta I_{CEO} - \frac{\Delta V_{B'E}\,h_{FE}}{R_E+R_B+r_{BB'}} + \frac{\Delta h_{FE}}{h_{FE}}\,I_C + \Delta R_B\,\frac{-I_C + h_{FE}\,(V_{CC}/R_1)}{R_E + R_B + r_{BB'}}\right] \quad \ldots\ldots\ldots\ldots (4)$$

where
$$S = \frac{R_E + R_B + r_{BB'}}{R_E\,(1 + h_{FE}) + R_B + r_{BB'}}$$

In the above equation, it has been implicitly assumed that $I_C \gg I_{CEO}$.

In order to keep $\Delta I_C = 0$, the required change in R_B is given by:

$$\Delta R_B = \frac{(R_E+R_B+r_{BB'})(-\Delta I_{CEO}-(\Delta h_{FE}/h_{FE}))+\Delta V_{B'E}h_{FE}}{-I_C + h_{FE}\,(V_{CC}/R_1)} \quad \ldots\ldots\ldots\ldots (5)$$

It is clear from equation (5) that if the variations of reverse saturation current, base-to-emitter voltage, and d.c. collector current with temperature are known for a given type of transistor, ΔR_B can be calculated.

The ratio of R_1 to R_2 depends upon the supply voltage, V_{CC}, and Thevenin's equivalent voltage, V_{BB}, and is given by:

$$R_1/R_2 = (V_{CC}/V_{BB}) - 1 \quad \ldots\ldots\ldots\ldots (6)$$

For a given operating point, the values of R_1 and R_2 can be chosen depending upon the input impedance requirement of the transistor circuit. For higher input impedance of the circuit and lower current drain from the supply, higher values of R_1 and R_2 are chosen.

Suppose that the operating point stabilization is to be done over the temperature range from T_X to T_Y, where subscripts X and Y denote the lower and upper limits, respectively. Now, choose a temperature T_O having a value nearer to the mean value of the above two temperature limits. Then, find out the components of the thermistor biasing network of the transistor, so that the change in collector current with temperature is completely balanced by the change in bias potential at temperatures T_X, T_O, and T_Y. Denote the required value of R_B, to do the above function, by R_{BX}, R_{BO}, and R_{BY}, respectively.

It is clear from equation (5) that the variation of ΔR_B over the temperature range from T_X to T_Y is not linear. At the higher temperature side, the variation of I_{CEO} will control the required value of ΔR_B. (The contribution of variations of reverse saturation current, base-to-emitter voltage, and d.c. collector current amplification factor in ΔR_B can be calculated separately from equation (5), if required).

Now, the values of R_{BX} and R_{BY} can be written as:

$$R_{BX} = R_{BO} + \Delta R_{BX} \quad \ldots\ldots\ldots\ldots (7)$$

$$R_{BY} = R_{BO} - \Delta R_{BY} \quad \ldots\ldots\ldots\ldots (8)$$

where ΔR_{BX} and ΔR_{BY} are the changes in R_{BO} calculated from equation (5).

So, the values of R_{2X}, R_{2O}, and R_{2Y} can be calculated from the following relations.

$$R_{2X} = \frac{R_1R_{BX}}{R_1 - R_{BX}} \quad \ldots\ldots\ldots\ldots (9)$$

$$R_{2O} = \frac{R_1R_{BO}}{R_1 - R_{BO}} \quad \ldots\ldots\ldots\ldots (10)$$

$$R_{2Y} = \frac{R_1R_{BY}}{R_1 - R_{BY}} \quad \ldots\ldots\ldots\ldots (11)$$

From Fig. 2, the values of R_{2X}, R_{2O}, and R_{2Y} are given by:

$$R_{2X} = \frac{R_A\,(R_S + R_{TX})}{R_A + R_S + R_{TX}} \quad \ldots\ldots\ldots\ldots (12)$$

$$R_{20} = \frac{R_A(R_S + R_{TO})}{R_A + R_S + R_{TO}} \quad \ldots\ldots\ldots\ldots\ldots (13)$$

$$R_{2Y} = \frac{R_A(R_S + R_{TY})}{R_A + R_S + R_{TY}} \quad \ldots\ldots\ldots\ldots\ldots (14)$$

Now, one has to choose the thermistor resistance, R_{TO}, at temperature T_O and find the values of series resistance, R_A, and shunt resistance, R_S, of Fig. 2, in order to satisfy the conditions imposed from equation (12) to (14). This can be done in the following way.

Eliminating R_A from equations (12) and (14) one obtains:

$$R_S^2 + bR_S + c - (d/a) = 0 \quad \ldots\ldots\ldots (15)$$

where $a = R_{2X} - R_{2Y}$

$b = R_{TX} + R_{TY}$

$c = R_{TX}R_{TY}$

$d = R_{2X}R_{2Y}(R_{TX} - R_{TY})$.

From equation (15), the value of R_S is given by:

$$R_S = \frac{-b \pm \sqrt{[b^2 - 4(c - (d/a))]}}{2} \quad \ldots\ldots\ldots (16)$$

In order to have a value of R_S equal to or greater than zero, one must have:

$$d/a \geqslant c$$

or: $(1/R_{TY}) - (1/R_{TX}) \geqslant (1/R_{2Y}) - (1/R_{2X}) \quad \ldots\ldots (17)$

So, before choosing the value of R_{TO} one must see that the condition given by equation (17) is satisfied for positive values of R_S. If this condition is not satisfied the balancing of collector current cannot be done at three temperature values.

When the proper thermistor is chosen, the value of R_S can be calculated from equation (16) and that of R_A from equation (13). Thus one can calculate analytically all values of the components of the thermistor biasing network for three point temperature compensation.

Design Considerations

The design will be considered of a thermistor biasing network for stabilizing the d.c. operating point of a Mullard type OC71 transistor over the temperature range 20 to 50°C. $V_{CC} = -6$V, $I_C = 1{\cdot}2$mA, $r_{BB'} = 0{\cdot}5$kΩ, $R_E = 1$kΩ (chosen).

For the above transistor, reverse saturation current doubles at approximately every 7·5°C increase in collector junction temperature; base-to-emitter voltage decreases at the rate of 2·5mV/°C increase in emitter junction temperature; d.c. collector current amplification factor approaches a maximum value depending upon the value of collector current, as shown in Fig. 1. Table 1 shows the various changes observed over the temperature range of 20 to 50°C for the same transistor.

Using the characteristics of the transistor and equation (6), $R_1/R_2 = 3{\cdot}3$. Choose $R_2 = 4{\cdot}7$kΩ, so, $R_1 \simeq 15$kΩ, and $R_{BO} = 3{\cdot}6$kΩ.

TABLE 1

Variation of reverse saturation current, base-to-emitter voltage and d.c. collector current amplification factor, for Mullard type OC71 transistor over the temperature range from 20 to 50°C

TEMPERATURE / VARIATIONS	$T_X = 20$°C	$T_O = 35$°C	$T_Y = 50$°C
ΔI_{CBO}	0·3mA	0	1·2mA
$\Delta V_{Bt'E}$	37·5mV	0	−37·5mV
Δh_{FE}	19	0	7

With the help of Table 1 and equation (5), $\Delta R_{BX} = 0{\cdot}221k\Omega$, and $\Delta R_{BY} = 0{\cdot}304k\Omega$.

From equations (7) and (8), $R_{BX} = 3{\cdot}82$kΩ, and $R_{BY} = 3{\cdot}29$kΩ.

From equations (9) and (11), $R_{2X} = 5{\cdot}13$kΩ, and $R_{2Y} = 4{\cdot}21$kΩ.

Now, choose the Siemens type K11 thermistor of value 20kΩ at 20°C. (This type of thermistor has 20 per cent tolerance and a temperature coefficient of −3·8 per cent/°C). For this type of thermistor, one finds that, in the present case, equation (17) is satisfied.

From equations (16) and (13), $R_S = 4{\cdot}4$kΩ and $R_A = 6{\cdot}5$kΩ, respectively.

Using the above calculated values in the thermistor biasing network of Fig. 2(a), the type OC71 transistor was tested for collector current variation over the temperature range from 20 to 50°C. Eight randomly selected transistors of the same type were put in the circuit one by one and their collector current variations over the temperature range was noted. The maximum collector current variation was found to be 0·08mA and the minimum was almost zero. The average value of the collector current variation was found to be 0·05mA, i.e. 4·1 per cent. Whereas by using a fixed value of resistance in place of R_{20} of Fig. 2(b), the average value of the collector current was found to be 0·28mA, i.e., 23 per cent.

During the above experiments, the junction temperature was kept approximately at ambient temperature. When a continuous current flows, as in actual cases, the limits of temperatures may accordingly be raised depending upon the dissipation and total thermal constant of transistor.

Conclusions

The analytical method developed above for the temperature stabilization of d.c. operating point of transistors by means of thermistor biasing network serves as an alternative to the graphical methods. For germanium transistors, the above method works well up to an ambient temperature of 50°C. The value of the intermediate temperature, T_O, need not be chosen in the centre of the temperature range. If it is chosen in such a way that the required variations in R_B about R_{BO} are equal, better stabilization can be achieved. The same procedure can be used for calculating the biasing network for stabilizing the operating point of silicon transistors.

Acknowledgment

The author wishes to express his indebtedness to Dr. G. N. Acharya, Head of Electronic Instrumentation Division, Central Electronics Engineering Research Institute, Pilani (Rajasthan), India, for his guidance throughout the work.

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NEW BOOKS

Thin Film Microelectronics

Edited by L. Holland. 284 pp. Med. 8vo. Chapman & Hall. 1965. Price 50s.

THE book is divided into six chapters by six different authors chosen for their acknowledged expertise in their field. One of the authors has been responsible for editing the whole to ensure uniformity of treatment. In this he has been reasonably successful, although some degree of overlap does still exist between chapters.

There are two natural parts in the book. The first is concerned primarily with physical properties and includes two chapters devoted to thin film passive components and active components respectively. The first chapter lists all the properties of the various materials used, both elemental and alloys, and examines the characteristics of resistors, capacitors, inductors and interconnexions achieved by these various means. This is a well documented chapter with comprehensive tables of physical properties which should prove useful to workers in this field. The second chapter is devoted to active devices and is necessarily more diffuse, dealing as it does with a subject still very much in the laboratory stage. A brief review of the various experimental processes in vogue is given, and the chapter concludes with the author's views on the likely future prospects for this technology.

This part of the book concludes with a chapter on semiconductor integrated circuits. This is expertly written and presented, with detailed descriptions of the various processes involved in the manufacture of silicon integrated circuits, including epitaxy, isolation, oxidation photolithography and the more complex processes such as sub-epitaxial diffusion, dielectric isolation, m.o.s.t. techniques, etc. The only apparent justification for the inclusion of this chapter in a book devoted to "Thin Film" microelectronics appears in the final two or three pages where compatible thin film/semiconductor technology is discussed.

The second part of the book is devoted entirely to the production problems of thin film circuits. A lengthy chapter, liberally illustrated, on the actual techniques and apparatus used for vacuum deposition is followed by a chapter on monitoring techniques where detailed discussions of the various methods developed for monitoring the deposition of thin films in terms of specified optical or electrical properties are given. The final chapter is devoted to the problem of the layout of thin film circuits and the masking and etching techniques, and this chapter, like the others, is well illustrated and extremely well documented.

The book as a whole is well produced. It would seem to be devoted more to the practice of thin film technology than to general microelectronics philosophies. This makes it difficult to understand the inclusion of the chapter on semiconductor integrated circuits, which although in itself eminently readable, appears totally out of place and in fact does break up the natural sequence of the book.

If this book is aimed at circuit designers in an attempt to assist in the implementation of the new techniques in future generations of equipment, then it would seem hard to justify the second part of the book with its wealth of manufacturing and processing details which in this context would seem irrelevant.

If the book is in fact aimed at the thin film specialist, it is quite possible that the reverse is true, and that the first half, and particularly the semiconductor chapter, is out of place. Such specialists would more likely turn to the more specialized tome on the subject written by the Editor of this present volume.

The book therefore may be of more restricted interest, aimed halfway between the two extremes stated above, and probably satisfying neither.

S. FORTE

Fields and Waves in Communication Electronics

By S. Ramo, J. R. Whinnery and T. Van Duzer. 754 pp. Med. 8vo. John Wiley & Sons. 1965. Price 102s.

FOR many years, Ramo and Whinnery's book 'Fields and Waves in Modern Radio' has been highly regarded by both students and professional engineers engaged in microwave and antenna work. This new book aims to update the earlier text. The treatment begins with a study of the elementary properties of resonant circuits, uniform and non-uniform transmission lines and the Smith chart. Static electric and magnetic fields are considered next and here the study is rather less formal than in many classic texts on electromagnetic theory, no prior knowledge of vectors is assumed and a number of excellent diagrams serve to make this chapter of real value to those confronted with electromagnetic theory for the first time. More detailed consideration of the solution of static field problems follows including transformation techniques and the method of separation of the variables. The latter topic affords the authors opportunity to reveal the fundamental properties of the essential mathematical functions appropriate to cylindrical and spherical problems.

The fundamentals of electromagnetism and Maxwell's equations are well discussed before circuit and impedance concepts are introduced. The chapter on propagation and reflection of plane waves includes propagation through an ionized gas and ends with brief but instructive comments on optics. There follows probably the best known chapters of the earlier work, guided electromagnetic waves and characteristics of common waveguides. The latter now includes a treatment of dielectric rod and slab guides, with photographs of fibre optics modes to illustrate the theory, also ridge and periodic waveguides are discussed. Propagation in anisotropic media has been allotted 40 pages and the authors consider dielectric anisotropy in crytsals, waves in plasmas and in ferrimagnetics; their approach is elementary, though they touch on the important fundamentals. The topic of resonant cavities is now embellished with a ten-page analysis of optical resonators for lasers. The section on microwave networks contains little new material and is confined entirely to analysis with no reference to synthesis. In the final chapter on radiation, valuable concepts of antenna theory are introduced and the radiation patterns of a number of basic antennas including microwave antennas evaluated. The concept of antenna impedance is given generous but justified space. Finally, reflecting the advent of lasers the diffraction grating is considered.

It is unlikely that any course in the United Kingdom follows the book from cover to cover, but it is of undoubted value for those studying electromagnetism, microwaves and antennas for the first time. The main question that arises in the mind of the reviewer is whether an omnibus edition is any longer the right way to treat a subject which has grown and is continuing to grow at a phenomenal rate. In many places in the book, more depth would have been valuable since only with this depth can the subject be placed in perspective. The lack of depth is felt most strongly with regard to the problems which are terse and, in many places, without even implied answers. In spite of these criticisms the book can be warmly recommended to those planning student libraries and it is to be expected that it will find its way on to the shelves of many practising engineers.

P. J. B. CLARRICOATS

Time-Domain Analysis and Design of Control Systems

By R. C. Dorf. 194 pp. Med. 8vo. Addison-Wesley Publishing Co. Inc. 1965. Price 68s.

NOWADAYS control theory is being applied in a wide range of fields, electrical, mechanical, industrial, biological and economic, and the problems of interest are often of great complexity. Increasingly the classical methods of control engineering are proving inadequate in obtaining useful solutions for the behaviour of complicated systems. Fortunately the development of the high-speed digital computer has provided a tool which can readily examine complex systems, provided the problems are suitably formulated. The methods of time-domain analysis supply the key to the expression of control problems in this way. It is against this background that these techniques are presented.

The first chapter is introductory, and the second chapter gives an introduction to the state space technique. A system is represented by a set of 'state variables' related to each other by a series of simultaneous first-order differential equations. Solutions as a function of time can be found for the state variables, provided the 'fundamental matrix', which is an expression of the dynamics of the system, is evaluated.

In chapter 3 some useful state space methods are given, for evaluating the 'transition matrix', which gives solutions for an arbitrary initial time, for linearization of non-linear systems, and for transformation to another co-ordinate system.

In chapter 4 system stability is considered by examining the sign of the time derivative of the energy (Liapunov's method), then this is extended to non-linear systems (Krasovski's method).

For a system to be controllable it is necessary to be able to observe all the state variables. This is not always possible in practice, and the implications of this are examined in chapter 5.

In chapter 6 and 7 the methods are extended to discrete-time, or sampled data systems, and in chapter 8 to non-linear systems.

In chapter 9 the design and synthesis of control systems is considered, and in chapter 10 methods for achieving optimal control are given.

This book gives a very clear exposition of the mathematics of time-domain analysis, and illustrates the methods by well chosen examples. It provides a very useful introduction to the subject.

G. D. BERGMAN

Active Network Synthesis

By Kendall L. Su. 369 pp. Royal 8vo. McGraw-Hill Book Co. 1965. Price 108s.

UNTIL recently the term 'active networks' was used mainly for oscillators or amplifiers, both of which need *active* devices (like transistors or tunnel diodes) in addition to passive elements (like resistors (R), inductors (L), capacitors (C)), in order to produce continuous oscillation or power amplification. In the last decade the meaning of the term has been extended to cover any networks, designed for signal shaping or frequency selection, which include active elements. For such networks the constraints governing passive and reciprocal (i.e. RLC) networks are decisively relaxed so that new methods of design and new types of circuits become available. Moreover the inclusion of active elements enables the designer to dispense with inductors and to obtain any required performance characteristic with active RC networks, which are realizable as thin film and monolithic integrated circuits.

The book by Prof. Su (of the Georgia Institute of Technology, U.S.A.) deals with this general type of active networks (and with some new types of amplifiers). Non-reciprocal elements (gyrators), although not necessarily active, are also considered, since they are usually realized with active components. The author "assumes that the students are well acquainted with such topics as Hurwitz polynomials, reactance functions, positive-real functions, cascade synthesis, synthesis of RC networks, and Darlington's theorems. A reasonable familiarity with matrix notation and operations is also essential . . . ".

In a short "Introduction" some basic matrix and network notations are defined. This is followed, in Chapter 2, by a review of various active elements (negative resistances, controlled sources (ideal amplifiers), negative impedance convertors (n.i.c.'s) and invertors) some non-reciprocal elements (gyrators, circulators) and the "pathological" one-ports "oport" or "nullator" (current $i = 0$ *and* voltage $v = 0$) and "support" or "norator" (both i and v arbitrary). Chapter 3 (49 pages) is entitled "Some basic properties of LLF (i.e. linear, lumped, finite) networks". In this chapter the basic theory (analysis) of active networks is discussed, at a high mathematical level, in terms of their energy functions (Hermitian forms). Necessary and sufficient conditions for functions to be realizable as immittance (impedance or admittance) functions of various types of active (and passive) networks are derived. Chapters 4 to 8 (180 pages) deal with the synthesis of a great variety of active networks; one-ports, two-ports and n-ports, using as active elements controlled sources, n.i.c.'s and negative resistances are discussed, and the important topics of stability and sensitivity are considered. Surprisingly, one important and powerful synthesis method is not explicitly mentioned, the utilization of operational amplifiers as differentiators, integrators and phase invertors, as used in analogue computing. The last chapter deals, in a very interesting way, with various types of negative-resistance amplifiers (transmission type, reflection type, etc.) using tunnel diodes as active devices. In two appendices various matrix theorems and scattering matrix and system matrix relationships are listed and briefly explained. Numerous examples, a bibliography (mainly U.S.A. sources), a name and a subject index are given.

This is an extremely useful book on the theory and synthesis—not on the engineering—of active networks, as it brings together and discusses in detail a great amount of material hitherto only available in periodicals. The treatment of the basic theoretical aspects occasionally suffers, in this reviewer's opinion, from too much compression, and it would seem desirable to expand it in future editions. It is certain that this book will be read and studied by everybody interested in the theory or in the practical design or active networks.

W. SARAGA

Man-Machine Engineering

By A. Chapanis. 134 pp. Demy 8vo. Tavistock Publications Ltd. 1966. Price 21s.

This book serves as an introduction to the subject of ergonomics and indicates the types of problems with which engineering psychologists are concerned together with the research methods used and some of the solutions obtained.

Colour Television Servicing Handbook

By W. Hartwich. 208 pp. Med. 8vo. Philips Technical Library. 1966. Price 58s.

In the first volume the fundamentals of colour television are described so that the subsequent understanding of receiver circuits can be followed. The first six chapters deal with the three colour theory of perception and with the transmission technique from the television camera to the final transmitter signal.

The remaining chapters deal with the decoding and the reproduction of the colour information in the receiver.

Connectors, Relays and Switches

By G. W. A. Dummer and N. E. Hyde. 229 pp. Demy vo. Sir Isaac Pitman & Sons Ltd. 1966. Price 50s.

The sixth volume in the Radio and Electronic Components Series deals with electronic connexions and considerable information is given on wrapped joints and crimping methods together with design data on relays, switches, plugs and sockets.

Integrated circuit connexions are also covered.

Microwave Transmission Circuits

Edited by G. L. Ragan. 716 pp. Demy 8vo. Constable. 1965. Price 24s.

This book is an unabridged and unaltered republication of the work first published by McGraw-Hill in 1948 as Volume 9 in the Massachusetts Institute of Technology Radiation Laboratory series.

Principles of Microwave Circuits

Edited by C. G. Montgomery, R. H. Dicke and E. M. Purcell. 486 pp. Demy 8vo. Constable. 1965. Price 18s.

This book is an unabridged and unaltered republication of the work first published by McGraw-Hill in 1948 as Volume 8 in the Massachusetts Institute of Technology Radiation Laboratory series.

Radar System Engineering

Edited by L. N. Ridenour. 748 pp. Demy 8vo. Constable. 1965. Price 24s.

This book is an unabridged and unaltered republication of the work first published by McGraw-Hill in 1947 as Volume 1 in the Massachusetts Institute of Technology Radiation Laboratory series.

Theory of Servomechanisms

Edited by H. M. James, N. B. Nichols and R. S. Phillips. 375 pp. Demy 8vo. Constable. 1965. Price 18s.

This book is an unabridged and unaltered republication of the work first published by McGraw-Hill in 1947 as Volume 25 in the Massachusetts Institute of Technology Radiation Laboratory series.

Electrical & Electronic Trader Year Book 1966

480 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 35s.

Principles of Aerial Design

By H. Page. 172 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 50s.

In this book, the author avoids the complications of advanced mathematics on the one hand and on the other the over simplication so that the treatment becomes purely descriptive.

The theory of aerials is developed from first principles and where approximations are necessary to avoid mathematical difficulties the reasons are given enabling a short introduction to be given to the theory of aerials while the practical forms of aerials used in the frequency range of 30kc/s to 30 000Mc/s are briefly described.

BBC Handbook 1966

259 pp. Crown 8vo. British Broadcasting Corporation. 1965. Price 7s. 6d.

An Introduction to Electronic Analogue Computers

By M. G. Hartley. 155 pp. Crown 8vo. Methuen & Co. Ltd. 1966. Price 10s. 6d.

This monograph provides an introduction to the subject for final engineering students as well as graduate engineers and scientists.

The features of computer and amplifier design are dealt with and while reference is not made to specific machines the treatment is indicative of modern practice.

The Glow Discharge and an Introduction to Plasma Physics

By F. Llewellyn-Jones. 211 pp. Crown 8vo. Methuen & Co. Ltd. 1966. Price 30s.

This monograph is intended for honours degree students of physics and electrical engineering and gives a full account of glow discharge initiation, through the spatio-temporal growth and development of space charge until the positive column in the steady state.

A number of practical applications is given.

Modern Electronic Components

By G. W. A. Dummer. 516 pp. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1966. Price 63s.

A revision of all data and information has been made in this second edition which includes a new chapter on Temperature, Humidity and Vibrating Categories of Components and another chapter dealing with changes in Component Characteristics with Time.

New Applications of Modern Magnetics

By G. R. Polgreen. 330 pp. Dem. 8voy MacDonald & Co. 1966. Price 60s.

Two main applications using ceramic permanent magnets are described in this book. The first deals with the re-design and improvement of existing machines such as electric motors, generators etc., while the second is concerned with novel developments such as magnetic suspension of travelling or rotating bodies.

A Vector Approach to Oscillators

By H. G. Booker. 149 pp. Demy 8vo. Academic Press. 1965. Price 19s. 6d.

High Vacuum Pumping Equipment

By B. D. Power. 412 pp. Med. 8vo. Iliffe Books Ltd. 1966. Price 80s.

A survey is given of the whole field of high vacuum pumping equipment with information on the correct choice of equipment and design of systems.

A description is given of oil sealed rotary pumps, oil and mercury diffusion pumps, vapour booster and ejector pumps, steam ejectors, mechanical boosters, molecular drag and turbomolecular pumps.

Semiconductor Counters for Nuclear Radiations

By G. Dearnaley and D. C. Northrop. 459 pp. Demy 8vo. E. & F. N. Spon. 1966. Price 8s.

The second edition of this book includes new developments such as the preparation and use of lithium drifted germanium devices for gamma-ray spectrometry.

Also included is a classified bibliography of some 600 associated papers.

Tables of Laplace Transforms

By W. D. Day. 36 pp. Demy 8vo. Iliffe Books Ltd. 1966. Price 5s.

Electronic Drafting and Design

By N. M. Raskhodoff. 594 pp. Med. 8vo. Prentice-Hall International. 1966. Price 96s.

This book is intended for the draughtsman who has a basic knowledge of engineering drawing and intends to apply it to electronic engineering. It covers components, graphical symbols, standards, printed circuits and so on based on American practice.

Materials Used in Semiconductor Devices

Edited by C. A. Hogarth. 243 pp. Med. 8vo. John Wiley & Sons. 1966. Price 105s.

This book, edited by C. A. Hogarth, contains chapters each written by separate authors and dealing with the following semiconductor materials: Germanium, Silicon, Selenium, Lead Sulphide, Selenide and Telluride, Indium Antimonide, Bismuth Telluride and Antimonides of Zinc and Cadmium.

Grundlegende Untersuchungen über die Möglichkeit der quantitativen Ausmessungen von Schallfeldern im Hinblick auf einen schalloptischen Analogrechner (Basic feasibility studies of the quantitative measurement of sound fields employing an acoustic-optical analogue computer)

By R. Pohlman, H. D. Bohnen, R. Brinkmann and D. Davidts. 55 pp. Royal 8vo. Westdeutscher Verlag, Cologne. 1965. Price DM.33S0.

An attempt is made to lay the foundations for an analogue computing method which allows indirect solution of wave equations for the case of two-dimensional propagation. The solution is termed indirect as no absolute value is determined; instead locations are obtained in which a mean value in respect of time derived from the solution is equal to a certain preselected constant. Thus the mean value in time of the energy distribution in space and the energy propagation in the wave field may be obtained. The studies referred particularly to ultrasonic wave fields.

Transmission Lines, Antennas and Wave Guides

By R. W. P. King, H. R. Mimno and A. H. Wing. 347 pp. Demy 8vo. Constable. 1965. Price 16s.

This book is a corrected and enlarged version of the work first published by McGraw-Hill in 1945.

LETTERS TO THE EDITOR

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

Pulse Generation After a Pulse Group

DEAR SIR,—During the design of a digital system the problem of transforming pulse trains, representing numbers, into corresponding positions in time was encountered. This is equivalent to designing a unit capable of producing one pulse after a pulse group (pulses equidistant in time) applied at its input (see Fig. 1). As this unit may prove useful in pulse group counting and other applications as well, the design considerations are explained in some detail.

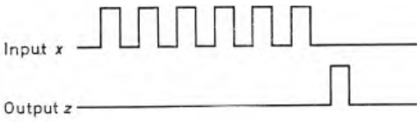


Fig. 1. Input and output pulses

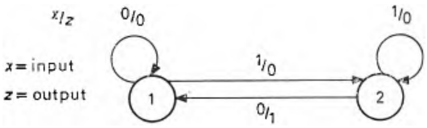


Fig. 2. State diagram

Fig. 2 shows the state diagram and Table 1 the state table of the proposed unit. It may be seen from the state table that the unit remains in position 1 when no pulses are applied at its input. When the first pulse appears the unit switches over to position 2 where it remains as long as pulses arrive at the input. When the pulse train is finished the unit switches back to position 1 giving one output pulse.

By state assignment the flow table (Table 2) is obtained from which the Boolean expressions for the output z and the next state w are obtained:

$$z = \bar{x}y \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$w = x \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

From the Boolean expressions the logic diagram of Fig. 3(a) is obtained, where D represents a unit delay. In agreement with the flow table, an R-S flip-flop with $R = \bar{x}$ and $S = x$ (see Fig. 3(b)) may substitute the delay element D, where now the invertor I_1 must be a pulse-logic invertor. Assuming that the clock producing the pulses in the pulse groups is available, the pulse-logic invertor may be realized as shown within the dotted

TABLE 1

y \ x	0	1	0	1
1	1	2	0	0
2	1	2	1	0
	w		z	

TABLE 2

y \ x	0	1	0	1
0	0	1	0	0
1	0	1	1	0
	w		z	

lines of Fig. 3(c) where I represents a level-logic invertor.

In Fig. 3(c) instead of the AND gate A of Fig. 3(a) and Fig. 3(b) a differentiating unit is used to produce the output pulse.

The logic diagram of Fig. 3(c) was realized with conventional transistor-resistor logic circuits and with a digit rate of 100kc/s. The photograph of Fig. 4 shows: (a) pulse groups applied at the input of the unit, (b) the corresponding output pulses, (c) the pulse-logic invertor output and (d) the R-S flip-flop output.

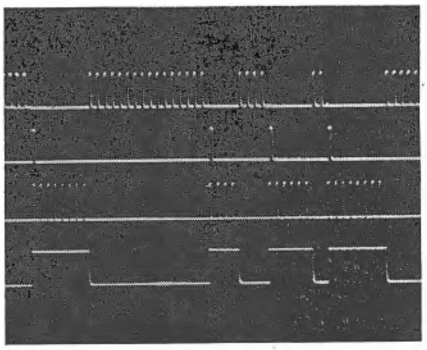


Fig. 4. Circuit waveforms
(a) Input (b) Output (c) Pulse-logic invertor output (d) R-S flip-flop output

The writers wish to thank Dr. J. Kontos for useful discussions and Mr. J. Karakatsanis for his help in the practical realization of the unit.

Yours faithfully,

G. PHILOKYPROU,
G. POLYCHRONAKIS,
Nuclear Research Centre "Demokeritus",
Athens,
Greece.

REFERENCE

HUMPHREY, W. S. Switching Circuits, pp. 212-254. (McGraw Hill, New York, 1958).

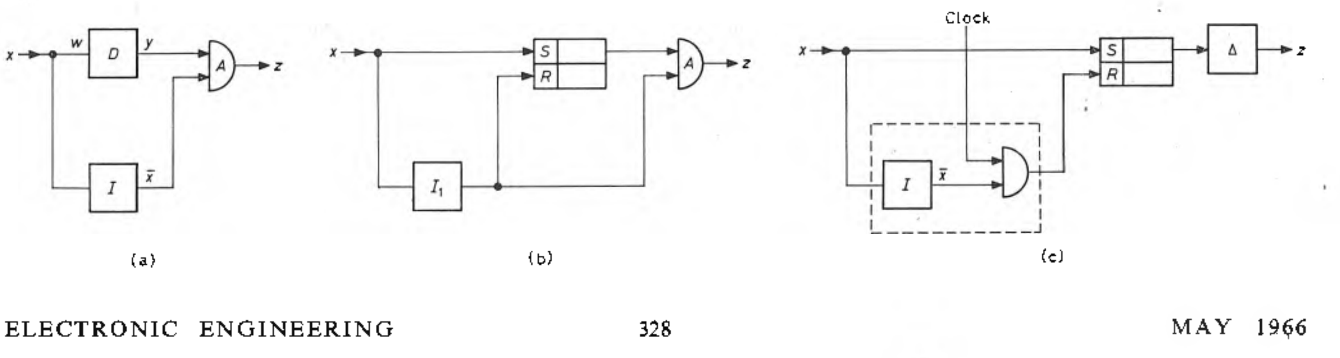


Fig. 3. Logic diagrams
(a) With delay element (b) With R-S flip-flop as delay element (c) Final logic diagram

The Achievement of High Overall Rejection in Difference Amplifiers

DEAR SIR,—Referring to the article by Mr. Salmons in the April, 1966, issue, I cannot agree with the statement that the effective cathode line impedance in Fig. 2 is the anode impedance of the device. It is shown below that the cathode line impedance is equal to $(r_a + R_k(\mu + 1))$, assuming that the grid of the constant current valve is connected to a d.c. potential (see Fig. 1).

In the majority of applications $(R_k(\mu + 1)) \gg r_a$, thus the cathode line impedance is mainly determined by R_k and the μ of the constant current valve.

If the common cathode line increases

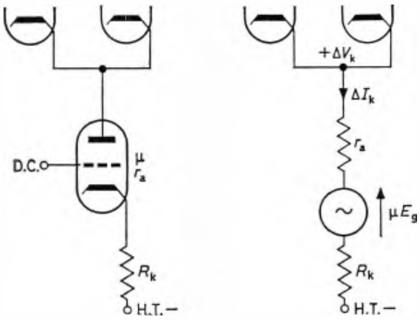


Fig. 1. Salmon's Fig. 2 and equivalent circuit

by a potential ΔV_k giving rise to a current ΔI_k, then:

$$\Delta V_k - \mu E_g = \Delta I_k (r_a + R_k)$$

But $E_g = \Delta I_k R_k$

Hence $\Delta V_k - \mu \Delta I_k R_k = \Delta I_k (r_a + R_k)$

$$\therefore \Delta V_k = \Delta I_k (r_a + R_k + \mu R_k)$$

Hence the cathode line impedance $= \Delta V_k / \Delta I_k = r_a + R_k (\mu + 1)$.

Yours faithfully,

R. W. J. BARKER,

Central Research Laboratory,
Edwards High Vacuum
International Ltd.,
Crawley,
Sussex.

The author replies:

DEAR SIR,—The 'constant current' device illustrated in Fig. 2 of my article is a cathode-loaded triode. The expression 'anode impedance of the device' therefore refers to the impedance seen on looking into the anode in Fig. 2, and not to the anode impedance of the valve alone.

I am grateful to Mr. Barker for pointing out this ambiguity and for emphasizing the overriding importance of the contribution of R_K to the total cathode line impedance. The expression '$r_a + (\mu + 1) R_K$' for the latter was in fact derived in the appendix of the D.I.C. Thesis (S. Salmons, Imperial College, London, 1964) referred to in my article, and was the one used in assessing the relative significance of the terms in the denominator of the expression for the rejection factor.

Yours faithfully,

S. SALMONS.

The University of Birmingham.

Effective Series Resistance of a Crystal

DEAR SIR,—During the study of a bridge crystal oscillator it was realized that the same scheme could be used to find the effective series resistance of the crystal. This brief note is submitted with the hope that this incidental realization might be of some interest to other readers.

A Wheatstone bridge configuration shown in Fig. 1, where one arm of the bridge is a crystal, can be made to oscillate by feeding across one of the diagonals from the output of an amplifier whose input in turn, is connected across the orthogonal diagonal of the bridge.

In this set-up, the crystal behaves like a series resonant circuit and the crystal appears as a simple resistor in the bridge. One of the arms R_3 of the bridge is made up of a thermistor and hence the circuit is level conscious. The presence of the thermistor also limits the amplifier output so that the amplifier is never saturated. If the amplifier gain were infinite, in principle, this set-up would oscillate at such an amplitude that the resistance of the thermistor (due to dissipation) becomes adjusted so that the bridge is at balance. Under these conditions, knowing the values of R_1, R_2 and R_3 the unknown resistance R_x can be calculated.

The resistance of the thermistor is a function of the output amplitude, its resistance can be evaluated from its characteristics, however this difficulty can be eliminated by an indirect method. If the oscillator output voltage V_1 and the voltage at one of the inputs V_2 is known, then R_x can be shown to be

$$R_x = \frac{V_1 - V_2}{V_2} R_2$$

It should be appreciated that in practice the amplifier is never at a perfect balance because of the finite gain in the amplifier. If necessary, of course, correction can be applied to take into account the finite gain of the amplifier.

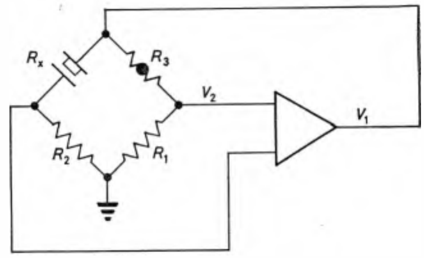


Fig. 1. The bridge circuit

The interest of Mr. P. H. W. Jacoby in this work and the assistance of Mr. R. G. Abbey in the trial measurements are thankfully acknowledged.

Yours faithfully,

R. KRISHNA,

Industrial and Research
Electronics Pty. Ltd.,
Sydney,
Australia.

REFERENCE

MEACHAM, L. A. The Bridge Stabilized Oscillator. *Proc. Inst. Radio Engrs.* (Oct. 1938).

Publications Received

DIAMOND RESEARCH 1965 is the second annual supplement to Industrial Diamond Review and consists of seven papers by authors concerned with various branches of research into diamond technology. It includes a paper on 'Electron Spin Resonance in Diamond' by Bleaney and Owen of the Clarendon Laboratory, Oxford.

Copies are available, price 5s. each from Industrial Diamond Information Bureau, Arundel House, 36-43, Kirby Street, London, E.C.1.

ELECTRICAL INSULATING MATERIALS is a twenty-page brochure recently issued by Attwater & Sons Ltd. It gives details of mica and micanites, slot insulations, laminates, tapes and sleevings, together with insulating papers stocked or manufactured by the company. This catalogue is available free on request to Attwater & Sons Ltd, P.O. Box 39, Preston, Lancashire.

ABRIDGED VALVE DATA—1966 is a quick reference brochure which is issued annually and is intended as a guide to EEV products current at the time of going to press.

The first part of the brochure gives abridged data for the range of valves and accessories produced at Chelmsford and Lincoln. The second part comprises an equivalents index and lists all EEV types, their applicable CV numbers and other manufacturers' types which they may replace.

Copies are available from The English Electric Valve Co. Ltd, Chelmsford, Essex.

POWER CONTROL FOR INDUSTRY is a new catalogue covering thyristor and diode stacks and assemblies with d.c. outputs up to hundreds of kilowatts.

Among other things the catalogue gives full details of heavy-duty diode rectifier stacks capable of providing up to a hundred kilowatts from 250 or 440V mains and also h.t. rectifier stacks having a typical rating of 6A at 18kV. Besides the Mullard range of heavy-duty stacks and assemblies intended for the most rigorous industrial conditions, information is also included on light-duty stacks and potted modules more suitable for electronic applications.

A page on thyristor drive modules shows the beginning of a full range which should simplify the application of thyristors to motor control.

In addition to cataloguing stacks and modules available from the company, this book gives detailed information on the thermal performance of the Mullard range of heatsinks, it provides charts for the selection of rectifier and thyristor stacks, an introductory chapter on many points that can arise when using aluminium extruded heatsinks, and data on h.t. modules and special purpose bridges.

Requests for copies of this comprehensive book should be made on company-headed notepaper to:— Mullard Ltd, Industrial Markets Division, Mullard House, Torrington Place, London, W.C.1.

INDUCTION BRAZING is a new technical data sheet on induction brazing that should be of interest to production engineers already using or contemplating the use of mass-production brazing techniques.

The data sheet explains why high frequency induction can provide the most rapid form of heating and is particularly useful in the brazing of large runs of ferrous components of relatively simple shape. Uniform and reproducible results can be obtained at high speed and heating cycles can be controlled by automatic timing devices.

The importance of correctly designed heating coils is stressed, and there are diagrams of most of the types in general use. Power considerations and the factors influencing choice of frequency are discussed. Advice is given on the selection of brazing alloys for particular joint clearances, and emphasis is laid on the advantages to be gained from correct alloy selection.

The new data sheet, 1100 : 143, is one of a series on low temperature silver brazing and is freely available from Johnson, Matthey and Co. Ltd, 73-83 Hatton Garden, London, E.C.1.

NEW EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

(Voir page 339 pour la traduction en français; Deutsche Übersetzung Seite 346)

PHASE-SENSITIVE DETECTOR

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustrated below)

Brookdeal Electronics Ltd are producing a new transistorized phase-sensitive detector which covers the frequency range 10c/s to 300kc/s and incorporates its own meter read-out while providing outputs suitable for driving a wide variety of pen recorders, digital voltmeters etc.

The high degree of linearity and low zero drift enable this instrument to be operated at very low signal levels in the presence of much larger noise voltages and it is particularly useful as an a.c.

detector in precision systems. Special attention has been paid to reducing setting-up operations to a minimum so that frequency range selection and fine adjustment of the reference voltage are eliminated.

The monitoring circuit, which includes a large scale meter with centre or left-hand zero, has comprehensive facilities which include switched time-constants ranging from 0·03 to 3sec, push-button attenuators, an optional filter for low frequency operation and zero suppression controls.

The unit is suitable for 19in G.P.O. rack mounting and can be supplied in an instrument case or in a dual unit case with suitable pre-amplifier.

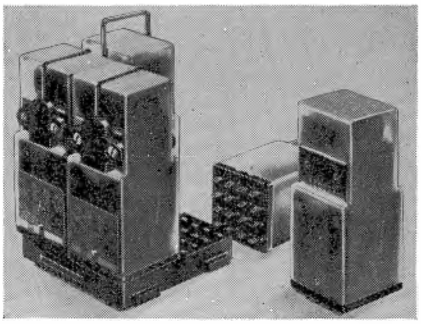
EE 93 751 *for further details*

RELAYS

Painton & Co. Ltd, Kingsthorpe, Northampton

(Illustrated above right)

Painton & Co. Ltd has entered into an exclusive licence agreement with Chauvin Arnoux, Paris, covering the manufacture of their entire range of OK relays in Great Britain. Painton has the exclusive selling and distribution rights for these relays, not only in Great Britain, but in Eire, Australia, New Zealand, Canada and South Africa.



Having a maximum current rating of 10A per contact with four changeover contacts and an extended mechanical life of several million operations, these relays cover a wide range of uses, models being obtainable for circuit breaking, time delay and logic applications and various other more specialized purposes.

New design techniques keep the relay in permanent adjustment and both the magnetic and mechanical circuits are extremely reliable. All variants of these relays are built into a standard transparent dust-proof case of compact dimensions, and both a.c. and d.c. versions are available. A full range of mounting accessories is available and base connexions are spade type, suitable for socket mounting or push-on type connexion.

EE 93 752 *for further details*

DIGITAL VOLTMETERS

Distributed by: Kynmore Engineering Co. Ltd, Buckingham Street, London, W.C.2

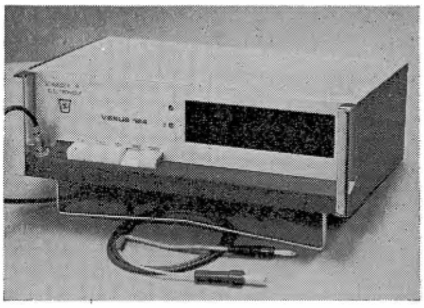
(Illustrated above right)

A new digital voltmeter of 0·1 per cent accuracy, modular construction and ergonomic styling is available from Kynmore Engineering Co. Ltd. Known as the Venus digital voltmeter, it is manufactured by the French Schneider company and is available in two standard models, each with a choice of refinements which, together with a range of ancillaries, make it suitable for high-quality measuring applications.

One standard model, the Venus 123, has three measuring ranges: $-9{\cdot}99$ to $+9{\cdot}99$V, $-99{\cdot}9$ to $+99{\cdot}9$V and -999 to $+999$V d.c. Automatic decimal point and polarity indication are provided, and accuracy is 5×10^{-4} of the full range reading ± 1 digit.

The other standard model, the Venus 124, contains a single additional digit to give overload indication. The maximum reading in each of three ranges is thus 10·00, 100·0 and 1 000V respectively, but higher values are indicated.

Allowable overload is approximately 25 per cent for both models, the 1 000V range excepted. Input impedance is 20MΩ



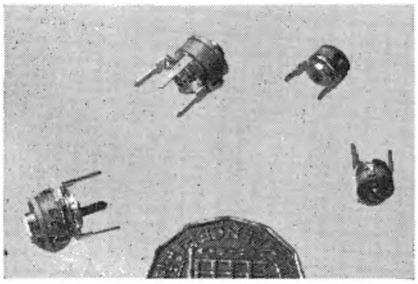
minimum, with 1 000MΩ optional on the 10V range. Operating temperature is $+5$ to $+45$°C nominal. The measurement is renewed automatically once every second.

Optional facilities include a printer output, and either model will operate with a Schneider printer (also available from Kynmore) or with an ancillary Schneider instrument (the 'Mars') which converts it into a digital multimeter. The Mars unit provides two lower d.c. ranges down to 100mV full scale, four a.c. ranges from 300mV to 300V r.m.s. full scale, as well as resistance ranges. Accuracy, sensitivity and input impedance depend on the particular range in use, but are comparable with those of the Venus alone.

The principle of operation is pulse-duration modulation, the Venus being fitted with a differential p.d.m. detector controlled by a quartz-crystal oscillator. Automatic polarity and voltage indication are achieved by internal logic, and the reading appears on cold-cathode display tubes. Range or calibration is selected by means of 'piano' keys, but the reference is set internally. Once set, it remains stable over very long periods after the initial warming-up period of ten minutes.

EE 93 753 *for further details*

ELECTRONIC ENGINEERING will occupy stand number N 816 in the National Hall, Olympia, at the I.E.A. Exhibition from 23 to 28 May, and visitors will be welcome.



CERAMIC DISK TRIMMING CAPACITORS

Steatite Insulations Ltd, Hagley House, Edgbaston, Birmingham, 16

(Illustrated above)

Steatite Insulations Ltd has made new additions to its range of subminiature ceramic disk trimming capacitors.

On the left of the illustration is shown the type 7 S-Triko 03 which has been designed to meet the exacting requirements of military and professional users. The trimmer conforms to MIL-C 81A and is available in the following capacitance ranges:

P100	2 to 3·5pF
N033	2·5 to 6pF
N075	3 to 10pF
N470	3·5 to 13pF
N750	4·5 to 20pF
N1500	7 to 35pF

Illustrated on the right is the type 5 S-Triko 03, a range of highly reliable ceramic disk trimmers available in the following capacitance ranges:

P100	1·5 to 2·5pF
N033	2·5 to 6pF
N075	3 to 8pF
N470	3·5 to 11pF
N750	4·5 to 15pF
N1500	6 to 22pF

EE 93 754 *for further details*

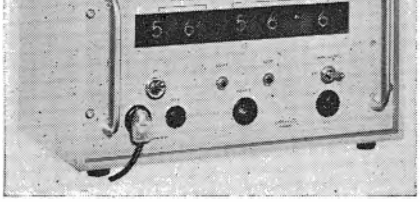
DIGITAL STOP CLOCK

Darang Electronics Ltd, Restmor Way, Hackbridge Road, Hackbridge, Surrey

(Illustrated below)

The type 661D is one of a new range of 'Digicron' electronic digital stop-clocks from Darang Electronics Ltd. The design is based on a system of low cost plug-in boards and uses extremely reliable cold cathode techniques. Many variations are possible to meet specific user requirements.

A five-digit in-line display is provided using gas-filled digital display tubes. The timing period covered is 59min 59·9sec in units of 0·1sec. On the standard model the mains supply frequency is used as the timing source, but provision is made for using an external timing source. A timing source employing a crystal in a temperature controlled oven can be incorporated at extra cost.



Stop and start is controlled by push-buttons on the front panel or by remote external contacts. The cable length to the remote contacts can exceed one mile.

Reset is automatic on operating the 'start' contacts or a number of time periods can be added and the total time displayed.

Power supplies required are 95 to 130V or 190 to 260V 50c/s.

EE 93 755 *for further details*

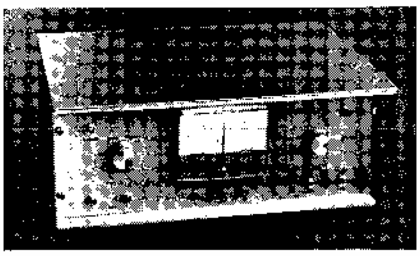
GALVANIC SKIN RESISTANCE METER

Short Brothers & Harland Ltd, Precision Engineering Division, Castlereagh, Belfast, N. Ireland

(Illustrated below)

The Short g.s.r. meter is a self-contained instrument powered from 230V a.c. mains. It enables the operator to measure:

(a) The skin resistance of a subject;
(b) The changes in resistance which occur in response to various stimuli, both anxiety provoking and neutral.



The skin electrode leads plug into two sockets on the front panel and a constant current of 10μA is passed through the skin path under test. The front panel meter shows the basal resistance and is not affected by variations in resistance. Three ranges are provided giving full-scale readings of 100kΩ, 200kΩ and 500kΩ. A front panel jack is used to connect a pen-recorder which records only the changes in resistance and shows zero when no change is taking place.

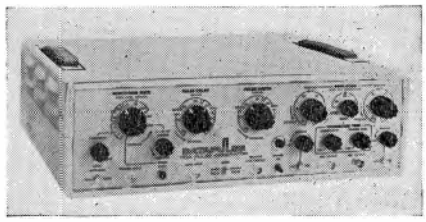
This separation of basal resistance and resistance changes is a great advantage over many other g.s.r. meters which require the operator to adjust potentiometers to back off the basal resistance in order to observe the changes. The only adjustment necessary with the Short instrument is a daily setting of the amplifier zero controls on the front panel.

EE 93 756 *for further details*

PULSE GENERATORS

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated above right)



Two new versions of the Datapulse 110 solid state pulse generator are now being manufactured in the U.K. by Dynamco Instruments Ltd. These are the 110A, which features improved rise-time and external triggering capabilities at no increase in price, and the 110P which is a programmed version of the 110A.

The 110A offers a minimum rise-time of 4·5nsec (4nsec typical) and external triggering from 1V. Repetition rate is variable from 4c/s to 40Mc/s. Linear transition times are separately variable. Simultaneous positive or negative outputs are available to 10V into 50Ω with up to 70dB attenuation in either single or double pulse modes. Pulse duration is variable from 10nsec to 5msec, pulse delay is variable from 10nsec advance to 50msec delay. Duty cycle is typically 85 per cent. Variable d.c. baseline control allows selection of complement output pulses at any level between $\pm$10V. Wide-range controls permit tolerance testing, facilitate repeatability of test results, and allow exact simulation of ramp and triangular waveforms for pulse inductance measurements and amplifier tests. Single pulse output permits trouble-shooting of logic circuits. Output states will not be damaged by any combination of panel control settings, short-circuits of any duration, or back voltages up to $\pm$10V. The unit is rated for operation with ambient temperatures from 0° to 50°C.

The 110P series programmed pulse generators allow remote control and/or automatic programming of all pulse parameters, including rise-time and d.c. baseline offset. High speed component test, such as test of integrated circuits, can be conducted through use with accessory equipment and card readers, punched paper tape, or magnetic disk programming.

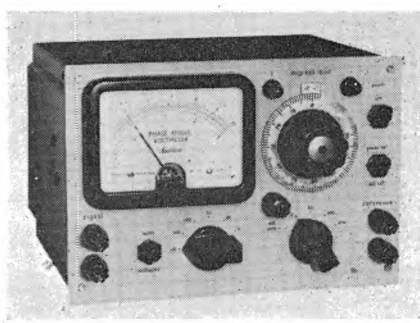
EE 93 757 *for further details*

PHASE-ANGLE VOLTMETER

Smiths Industries Ltd, Aviation Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

A new compact solid-state 400c/s phase-angle voltmeter incorporating many



useful features is now available from the Aviation Division of Smiths Industries Ltd. Measuring 8¼in long by 6in high by 8in deep, the instrument is accurate to ±3 per cent in amplitude and ±1 degree in phase.

The instrument can be used to present either the resolved components or the phase and modulus of the input, and sensitivity in 12 ranges is available between 300V and 1mV, an a.g.c. loop making manual level adjustment unnecessary. The single setting up control requires adjustment only when there is a change of signal frequency. Both the signal and reference inputs are transformer coupled, giving the high common-mode rejection ratio ($10^5:1$) required for valid bridge and differential measurements. Filters are included in both signal and reference channels to provide the necessary degree of harmonic rejection.

A 10 : 1 overload factor has been provided so that minor components can be measured in the presence of a large total signal. A warning lamp indicates when this factor has been exceeded.

The instrument will operate from any power supply between 100V and 250V, 50 to 400c/s. No adjustment is required when changing voltage or frequency—a feature which obviates the likelihood of mistakes where more than one kind of supply is available.

EE 93 758 *for further details*

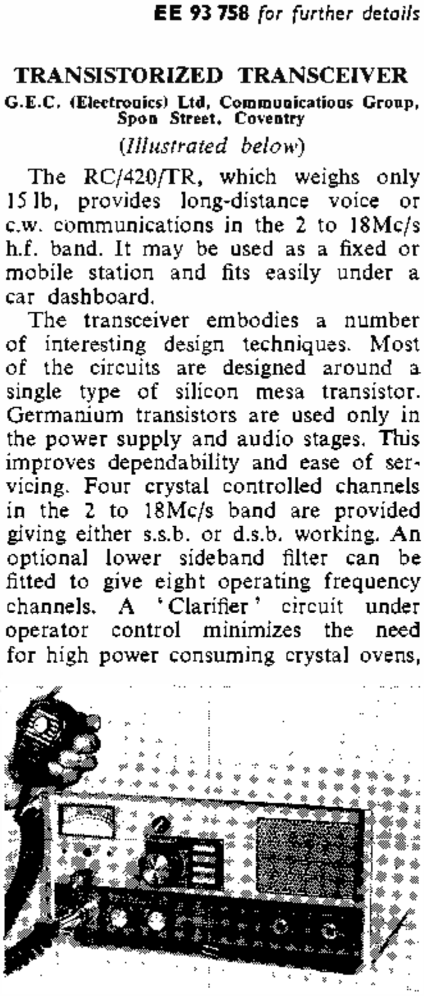
TRANSISTORIZED TRANSCEIVER

G.E.C. (Electronics) Ltd, Communications Group, Spon Street, Coventry

(Illustrated below)

The RC/420/TR, which weighs only 15 lb, provides long-distance voice or c.w. communications in the 2 to 18Mc/s h.f. band. It may be used as a fixed or mobile station and fits easily under a car dashboard.

The transceiver embodies a number of interesting design techniques. Most of the circuits are designed around a single type of silicon mesa transistor. Germanium transistors are used only in the power supply and audio stages. This improves dependability and ease of servicing. Four crystal controlled channels in the 2 to 18Mc/s band are provided giving either s.s.b. or d.s.b. working. An optional lower sideband filter can be fitted to give eight operating frequency channels. A 'Clarifier' circuit under operator control minimizes the need for high power consuming crystal ovens, thus cutting battery drain. All circuit boards and key parts are treated against fungoid growth.



The unit is supplied complete with a dynamic microphone. A wide range of accessories is available including a 'phone patch' for direct communication to telephone lines; a 110/220V a.c. power supply; an aerial tuner for multiple frequency use and a mounting rack and aerial installation for mobile applications. A conventional telephone handset and Teletype terminal and printer equipment can be supplied.

EE 93 759 *for further details*

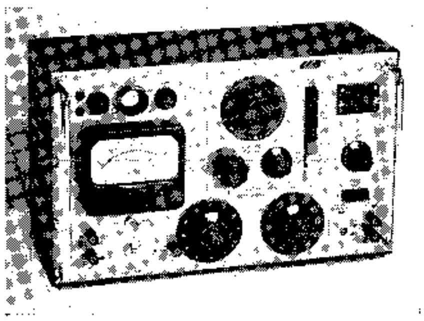
DISTORTION-DECIBEL-MILLIVOLTMETER

Laboratoire Electro-Acoustique, 5 Rue Jules Parent, Rueil (S & O), France

(Illustrated below)

The 'Qualiscope' is a universal type precision instrument combining in a single unit means for measuring the principal properties of a circuit. The parameters measured are: non-linear distortion (per cent), voltage (mV), voltage level (dB) and noise level.

The distortion meter covers fundamental frequencies from 10c/s to 100kc/s



and harmonic frequencies from 10c/s to 250kc/s, the maximum sensitivity being 0·1 per cent f.s.d.

The voltmeter-decibel-meter covers the frequency range of 2c/s to 400kc/s, the frequency response being within ±0·2dB from 20c/s to 100kc/s and ±2dB from 2c/s to 400kc/s. The voltage range is 1mV to 300V f.s.d. An additional pre-amplifier (gain +20dB) raises the sensitivity to 100μV f.s.d.

The instrument includes a built-in c.r.t. for the display of Lissajous figures.

EE 93 760 *for further details*

AUTOMATIC SCALER

L'Electronique Appliquée, 96-100 Rue Maurice Arnoux, Montrouge (Seine), France

(Illustrated above right)

This transistorized scaler has a display of six 'Nixie' tubes and a six-digit electromechanical register with push-button reset. The mode of operation is selected by push-buttons on the front panel.

The scaler will count pulses between an adjustable threshold of 0·1V to 10V, the input resolution being 1μsec. The input impedance for positive pulses is 10kΩ and for negative pulses 2·5kΩ.

For frequency measurement the counting time is 1sec and the read-out time the following 2sec.

The scaler can be stopped automatically after preselected counts of any number up to $n \cdot 10^p$; where n is adjustable from 1 to 9 and p is adjustable from 0 to 6. Alternatively the scaler can be stopped automatically after preselected times of 1 to 10^6sec. If a time and a count are both preselected, the scaler will stop at whichever is reached first.

The frequency standard is provided by a 100kc/s quartz crystal.

An output is provided for operating a printer.

EE 93 761 *for further details*

SERVO MOTOR

Servomex Controls Ltd, Crowborough, Sussex

(Illustrated on page 333)

This new small 50c/s two-phase servo motor with tachometer generator and reduction gearbox is of simple construction and exceptionally low cost, having been developed instead of the conventional type of servo motor which, being designed principally for military uses, has an outstandingly high performance and a correspondingly high cost.

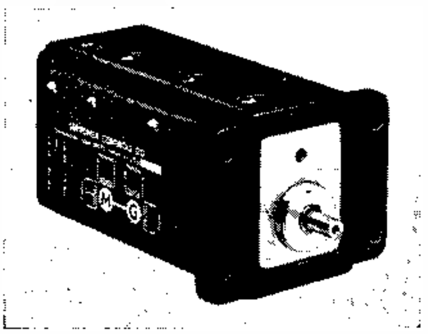
The materials and workmanship of the new Servomex motor are, however, of the highest standard and it is suitable for inclusion in instrument servo-mechanisms of the best quality.

The motor is of the four-pole type with skewed rotor bars and an elementary drag cup type tacho generator. Two models are available, SM 95 and SH 135, with different windings and bearings, but with basically the same performance. The output shaft of model SM 95 runs in a porous bronze bearing impregnated with silicone lubricant whereas model SM 135 has an output shaft running on a precision ball race and has an accurately machined spigot.

The coils are vacuum impregnated in accordance with Specification RCS 214. The gearbox, having a reduction ratio of 40 : 1, uses nylon and steel gears and incorporates a slipping clutch set to slip at approximately 40 to 55 oz in. This has been provided to prevent any damage or impact.

The output shaft has a maximum speed with no load of 60rev/min, the stalled torque being 30 oz in. Both models are built in an aluminium alloy casting and are suitable for an operating temperature range of −10° to +70°C.

The SM 95 is designed to be driven from an amplifier with a 90° phase change in it, both the motor reference



and generator excitation windings being fed from 6·3V a.c. of zero phase. The motor reference winding for the SM 135 is designed to be supplied from 230V r.m.s. through a capacitance of 1μF to 1·4μF. Thus the current in the winding leads the supply voltage by 90°, permitting the control windings to be energized from an amplifier of zero phase shift.

The dimensions are 3in by 4½in by 2⅜in and the weight is 2·19 lb.

EE 93 762 *for further details*

CERAMIC DISK CAPACITORS

Centralab Ltd, Northway House, Great North Road, London, N.20

This is a new range of low cost ceramic disk capacitors having quality performance and size advantage. The three capacitance values are 0·01, 0·02 and 0·05μF in disk diameters of 0·297, 0·359 and 0·516in respectively, and a thickness not exceeding 0·125in.

The working voltage of the range is 50V d.c. All sizes have a lead spacing of 0·25in except the largest, which is 0·375in, the leads being supplied either straight or terminated with a kink and cut for use in printed boards. All leads are solder coated for ease of assembly.

Each unit in the range is supplied at a tolerance value of either +80 per cent, −20 per cent or ±20 per cent.

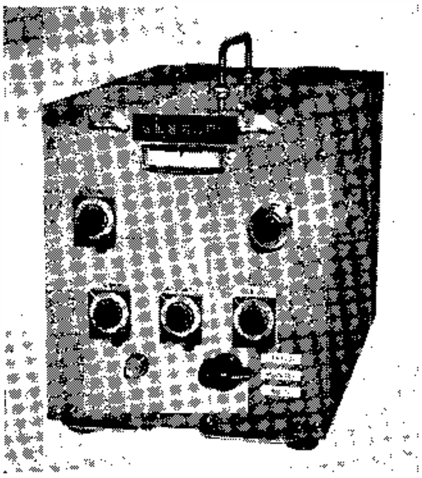
EE 93 763 *for further details*

TEMPERATURE CONTROLLER

Société D'Etudes et Recherches Electroniques, 15 Rue Villebois-Mareuil, 92-Colombes (Seine), France

(Illustrated below)

This new SERELEC temperature controller combines proportional, derivative



and integral control and will maintain close regulation, typically 0·2°C/500°C, despite fluctuations of mains voltage, ambient temperature or other external disturbances.

It comprises three magnetic amplifier stages with two thyristors for control of the output.

EE 93 764 *for further details*

ELECTRONICALLY TUNABLE YIG FILTERS

Microwave Associates International Inc.; European Office: Cradock Road, Luton, Bedfordshire

(Illustrated below)

A new range of electronically tunable YIG filters in the frequency range from 125 to 12 000Mc/s has been introduced by Microwave Associates. These filters have a very low insertion loss, good off-band rejection, and are free from spurious response.

Weighing only 1·8 lb they have a

maximum v.s.w.r. of 1·5 and meet all the requirements of MIL-E-5400, Class II.

The filters are highly polished, single crystalline spheres of yttrium iron garnet for use as gyromagnetic tuning elements.

Driver amplifiers are available, which, by means of a 0 to 10V signal from a high impedance source, drive the filters from the bottom to the top of their respective ranges. Power supply and temperature control units can also be supplied where necessary.

These units are designed to provide preselection and post selection for any type of receiver requiring rapid scan and/or automatic signal selection or signal processing.

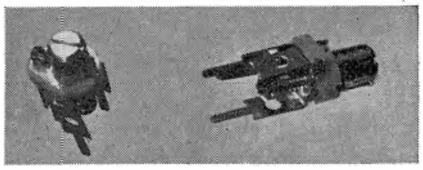
EE 93 765 *for further details*

TRIMMER CAPACITORS

Steatite Insulations Ltd, Hagley House, Hagley Road, Edgbaston, Birmingham, 16

(Illustrated above right)

A recent addition to Steatite Insulations' range of trimmer capacitors is a screw trimmer type 1250b. Designed for printed circuit mounting, it is available with a maximum capacitance of 10pF or 20pF; the minimum capacitance in either case being 2pF. The capacitance law is linear over a tuning angle of



6 × 360°. The power factor at 1Mc/s is better than 0·002 and the working voltage is 250V d.c. The overall size is 15·5mm by 7·5mm.

EE 93 766 *for further details*

D.C. AMPLIFIER

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Illustrated below)

S.E. Laboratories (Engineering) Ltd has introduced a fully floating differential input, high gain, wide band, low noise d.c. amplifier S.E.4200 for use with oscillograph and magnetic data recorders, data acquisition systems, etc.

Design of the S.E. 4200 makes it suitable for amplifying very low level signals, such as those derived from thermocouples, strain gauges and similar devices.

A feature is the high stability normally only achieved with chopper stabilized amplifiers. No chopper is used in this amplifier. Advantages are reduced noise and complete freedom from intermodulation distortion.

Another feature is high common mode rejection—greater than 120dB—and it is unaffected by input lead asymmetry of up to 1kΩ. Common mode voltages can be as high as 160V r.m.s. Either input lead can be grounded without affecting output zero.

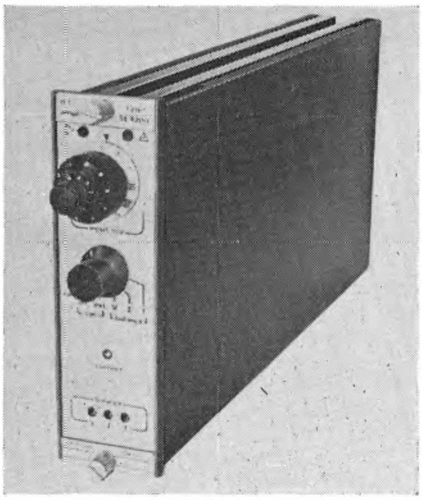
A ten-step attenuator has an overall ratio of 1 000 : 1. Additional intermediate 2·5 : 1 vernier gain control is given by a 10-turn potentiometer fitted with a lockable 1 000° division dial.

Model S.E. 4200 can be supplied with a number of options:

Output ±1·4V at ±8mA.
±10V at 100mA or 1mA.

Gain of the amplifier is 1 400 or 10 000 respectively, i.e. full output at signal input of 1mV.

Overall stability and accuracy is better



than $\pm 0{\cdot}1$ per cent of f.s.d. on attenuator setting of 10mV. Frequency response of the amplifier system is better than ± 10 per cent in the range of 0 to 10kc/s. Noise is better than 4μV peak-to-peak referred to input.

Additional options are a built-in calibration system which can be remotely operated or linked into data loggers or computers.

Each amplifier module measures 2in by $6\frac{1}{2}$in high by approximately 13in deep, and up to eight can be housed in a 19in rack-mounted cabinet.

EE 93 767 *for further details*

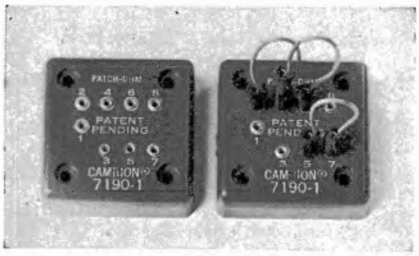
RESISTANCE PATCHBOARD

Cambion Electronics Products Ltd, Castleton, Nr. Sheffield, Yorkshire

(Illustrated below)

The Cambion 'Patch-Ohm' is a miniature resistance patchboard, measuring approximately $1\frac{1}{4}$in square, which provides 56 standard 5 per cent resistance values ranging from 51Ω to $9{\cdot}5$kΩ.

Fabrication of the 'Patch-Ohm' resistor is accomplished by advanced micro-electronic techniques. The device consists of a high reliability resistor thick-film network yielding standard resistors with 5 per cent tolerances developed by Microtek Electronics (MTE). The thick-film resistor network is glass passivated



by a new process. The temperature coefficient of the 'Patch-Ohm' resistor material is 200 in 10^6 from $-80°$C to $+150°$C. The power rating for the thick-film resistor materials is 15W/in^2.

The various resistance values are obtained by different patching between the eight sockets which accept Cambion 0·04in diameter 'piggy-back' patch cords.

EE 93 768 *for further details*

COMPONENT TESTER

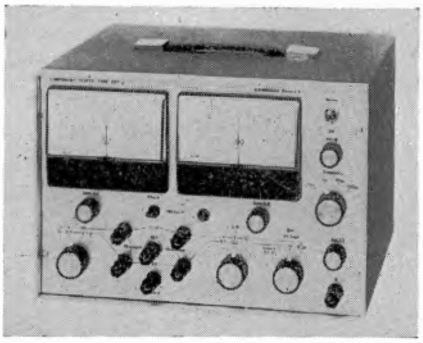
Distributed by: Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated above right)

The Instrument and Digital Measurement Division of Dawe Instruments Ltd, as sole U.K. agents for A/S Danbridge of Denmark, are now able to supply the *RCL* component tester type CPT2 which is manufactured by this Danish company.

The instrument provides a direct reading comparison bridge for comparing impedances and phase angles over a wide range of impedance and frequency. Special features of the instrument are its high accuracy and stability together with ease of operation.

The *RC* oscillator provides power supply frequencies of 100c/s, 1kc/s,



10kc/s and 100kc/s to the bridge circuit and facilities are provided for direct reading on two separate meters.

Typical applications are acceptance, production and grading tests on components and circuit parts, measurements on inductors, capacitors and electrolytic capacitors with superposed d.c. The instrument is also suitable for testing and adjusting ganged components.

Laboratory applications include the measurement of long time stability, temperature and humidity influence, high accuracy impedance comparison and phase measurements.

Impedance range is 3Ω to 5MΩ on 0·3V range, 100Ω to 5MΩ on 3V range. High impedance values up to 50MΩ may be measured to the same accuracy by calibration adjustment.

EE 93 769 *for further details*

PHOTO RELAY UNITS

O.P. Instruments Ltd, 246 Sutton Common Road, Sutton, Surrey

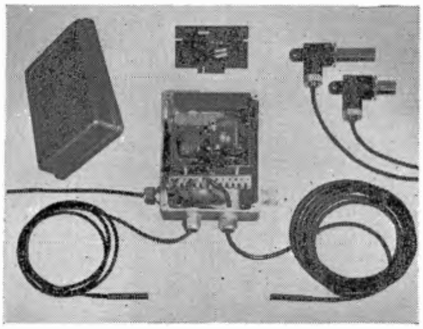
(Illustrated below)

In this new range of photo-relay units and heads all units are contained in weatherproof, cast aluminium cases. Plug in, replaceable circuit boards and relays are used throughout. Most systems can be fitted with a lamp failure alarm unit which will give warning in the event of lamp or lamp supply failure.

Apart from normal on/off operations, a variety of circuit panels can be fitted giving delay before relay operation or controlled, maintained relay closure.

Most units can be offered for normal light on/light off operation or for operation by a momentary light pulse. Circuits can also be arranged for operation by two photocells, light on one cell closes the relay, light on the other cell opens the relay.

A variety of cell and lamp heads are available. Miniature heads can, under



favourable conditions, detect objects as small as 1mm in diameter.

The units are normally arranged for mains operation, but portable battery operated systems can be supplied.

EE 93 770 *for further details*

STABILIZED E.H.T. POWER SUPPLY

Derritron Instruments Ltd, Parklands, Caincross, Stroud, Gloucestershire

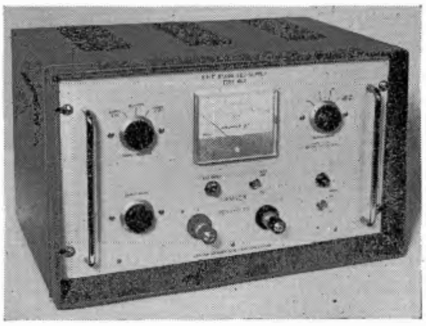
(Illustrated below)

Derritron Instruments Ltd, is offering an encased e.h.t. power supply at what is claimed to be the lowest price yet achieved for a unit of such range and quality.

The power supply, catalogue number E.0411, delivers from 500V to 3kV d.c. continuously variable, at up to 10mA. The stabilization factor is 2 000:1 and the ripple limited to 17mV peak-to-peak at full load. Output impedance is not greater than 100Ω.

The 0411 has been designed as a general laboratory supply and is built on a standard 19in front panel. It may be removed from its case and mounted in a rack if required. The output is completely isolated and is switched to provide either a positive or negative supply.

Stringent safety precautions have been



taken to ensure that all sockets, switches and potentiometers are isolated from the chassis.

EE 93 771 *for further details*

HIGH GAIN DIFFERENTIAL AMPLIFIER

Tektronix U.K. Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

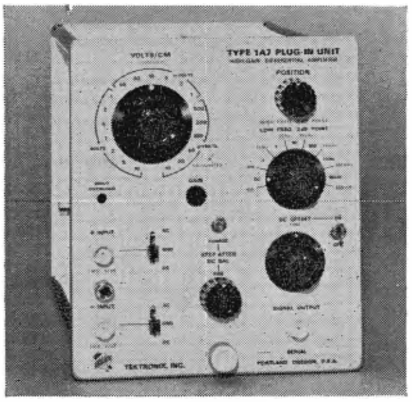
(Illustrated on page 335)

Simplified d.c. balancing procedure, wide range sensitivity and versatile frequency response selection make the type 1A7 a useful general purpose, high gain plug-in unit for use with Tektronix 530, 540, 550 and 580 (a type 81 adapter is required), series oscilloscopes.

The total calibrated sensitivity range is 10μV/cm to 10V/cm, with a d.c. to 500kc/s bandwidth availability at all sensitivities.

Selection of the amplifier bandwidth can be made by using two front panel switches providing 8 high frequency and 7 low frequency 3dB points.

Input noise is $3{\cdot}3\mu$V r.m.s. maximum (not including drift) and d.c. drift is no more than 200μV/h averaged over 10h



with ambient temperature and line voltage constant.

The type 1A7 can reject common-mode signals with amplitudes up to 20V peak-to-peak and frequencies up to 100kc/s, with a common-mode rejection ratio of 50 000 : 1.

Other features include a 300mV d.c. differential offset capability, a differential overload light and a signal output socket.

EE 93 772 *for further details*

CIRCUIT BOARD

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

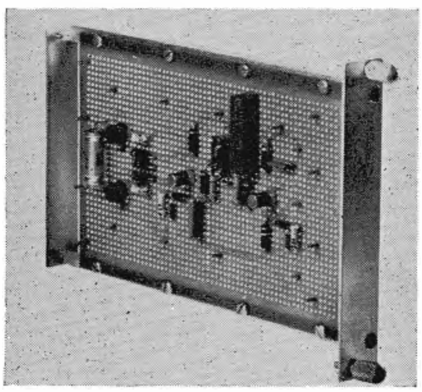
(Illustrated below)

A.P.T. Electronic Industries Ltd, announce an addition to their range of components for the MINAR modular chassis system.

The Minar circuit board C.B.1 is a phenolic insulated board or plate, measuring 6·4in × 4·25in, and is punched with an overall pattern of holes on the internationally standard 0·1in grid. Any electronic circuit may be built on the board with little or no drilling, since all wire ended components will mount directly on to the board and can be interconnected by their own wires on the reverse side. They may also be wired on to 'Lektrokit' soldering pins, which 'push fit' into the holes providing double ended points wherever required.

The circuit board is designed to mount on two tie rod supports behind the individual module front panels, each module being capable of carrying one or two boards.

The MINAR modular chassis construction system, derived from MINiature Apparatus Rack, was designed by the British Broadcasting Corporation Research Department for use in the construction of experimental stereo and colour television apparatus, and is manufactured and distributed by A.P.T. as a neat, durable and versatile chassis system for experimental circuit construction.



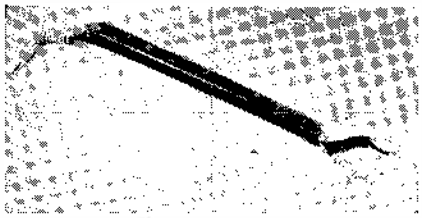
EE 93 773 *for further details*

TELEVISION DELAY LINE

Lexor Electronics Ltd, 25-31 Allesley Road, Coventry

(Illustrated below)

Anticipating the coming demands for specialized components for colour television, LEXOR has developed a low-cost delay line having the function of a low-pass filter.

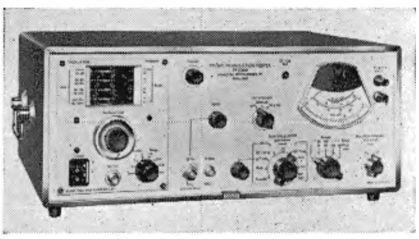


The application is as a filter for the phase correction of the luminance channel with respect to the chrominance channels. Exact parameters would depend upon the type of colour system, and the circuit technique employed.

A typical unit, as illustrated, would have a delay of 800nsec at an impedance of 1·8Ω and a cut-off frequency of 2·5Mc/s with linear phase response. Dimensions are 6in long by $\frac{3}{8}$in diameter and the finish is by plastic encapsulation.

If ordered in sufficient quantity, the price of these components could be extremely low.

EE 93 774 *for further details*



A.M.-F.M. MODULATION METER

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated above)

An entirely new, fully transistorized general-purpose modulation meter is announced by Marconi Instruments Ltd. The instrument, type TF 2300, is designed to meet all requirements in the fixed and mobile point-to-point communication field, and most of those in broadcasting and telemetry.

The carrier frequency range is 4Mc/s to 1 000Mc/s. Five deviation ranges, from 5kc/s to 500kc/s, are provided with modulation frequency range from 30c/s to 150kc/s. For measurements in the audio band, the modulation frequency range may be restricted to 15kc/s by means of a panel control.

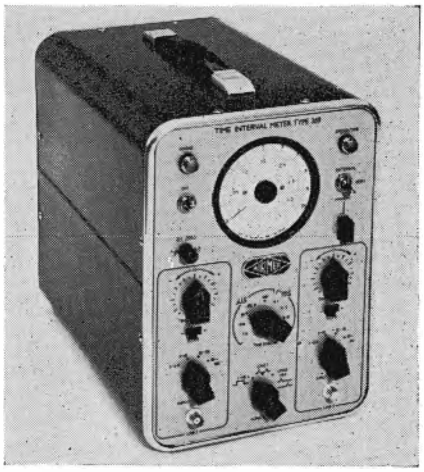
Two ranges of a.m. depth measurement are provided, 0 to 30 per cent and 0 to 100 per cent at fundamental modulation frequencies in the range 30c/s to 15kc/s. The modulation frequency range extends to 50kc/s for distortion measurements.

A demodulated signal output is available at a level of 1mW into 600Ω for meter f.s.d., both for f.m. and a.m. signals. A standard 50μsec or 75μsec de-emphasis filter may be switched into the demodulated signal output circuit for measurements on transmitters using pre-emphasis.

The local oscillator incorporates facilities for crystal controlling the frequency to eliminate local oscillator noise, and to permit rapid production testing at up to three predetermined carrier frequencies. An external local oscillator input facility is also provided. A built-in crystal calibrating source enables the deviation indication to be standardized.

The instrument will operate on normal mains voltages, or on a nominal 24V d.c. positive earth supply.

EE 93 775 *for further details*



TIME INTERVAL METER

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustrated above)

The time interval meter, type 369, is an inexpensive and fully-transistorized equipment for measuring very short increments of time in the range 100nsec to 30msec.

Essentially a very accurate electronic stop-clock employing counting techniques in conjunction with stable internal oscillators, the instrument operates from positive and/or negative going steps or pulses and can be set to trigger at any level from 1V to 100V; either one or two-line working can be used.

This combination of facilities enables measurements to be made of pulse width, pulse spacing, and rise or fall times, in addition to measuring the time interval between pulses on two different lines.

Indication is on an open-scale 5in meter, the reading being held indefinitely until reset.

EE 93 776 *for further details*

SHORT NEWS ITEMS

A Conference on Electrical Networks is to be held in the Department of Electrical Engineering at the University of Newcastle on 13 to 15 September this year. The conference is sponsored jointly by the University of Newcastle upon Tyne, the Ministry of Technology, the Institution of Electrical Engineers and the Institution of Electronic and Radio Engineers.

Further details, including registration forms, are obtainable from the Conference Secretary, Ministry of Technology, Northern Regional Office, Wellbar House, Gallowgate, Newcastle upon Tyne 1.

The Fourth International Measurement Congress (IMEKO IV) is to be held at Warsaw, Poland, on 3 to 8 July 1967, and papers will be presented on the following subjects:

(1) New methods of measurements, research on basic and derived standards.
(2) Instrument design, manufacturing technology and economics.
(3) Instrument application techniques, aspects of industrial applications and public services.

Offers of papers from the U.K., which should include a synopsis in triplicate not exceeding 500 words in length, should be forwarded not later than 1 June this year to the Secretary, The Society of Instrument Technology, 20 Peel Street, London, W.8.

The Institution of Engineering Inspection, in collaboration with the Materials and Testing Group of The Institute of Physics and The Physical Society, is arranging a conference on "The Economics of Automated Materials Testing" to be held at Imperial College of Science and Technology, London, on 4 to 6 July this year.

The conference will deal with the following subjects:

(1) Economics of automated inspection.
(2) Automatic inspection.
(3) Inspection as an integral part of automated production.

Advanced registration for attendance at this meeting will be necessary and further details and application forms may be obtained from The Secretary, The Institution of Engineering Inspection, 616 Grand Building, Trafalgar Square, London, W.C.2.

The Television Society has awarded the "Electronic Engineering" premium to E. J. Gargini of Rediffusion Research Ltd for his paper entitled, "Colour Television by Wire."

The Radio and Electronic Component Manufacturers' Federation Information Bureau on United Kingdom sources of supply of components, accessories and materials for the radio, electronic and allied industries is now available for use by engineers in the United Kingdom. Users of electronic components who wish to obtain these requirements from U.K. sources are asked to contact the RECMF Secretariat, 6 Hanover Street, London, W.1 (Telephone MAYfair 2472).

The Electrical Engineering Division of the School of Engineering, the University College of Swansea is holding a one week course on Field Effect Electronics on 4 to 9 September this year. The object is to familiarize practising circuit engineers and others, of the present state of f.e.t. development and to show how these can be used in a wide range of applications.

The course will be a residential one, all members being accommodated in a hall of residence on the college campus. The inclusive fee will be £25.

Further details are obtainable from the Course Secretary, Dr. W. G. Townsend, School of Engineering, University College of Swansea, Singleton Park, Swansea.

The new Post Office high frequency radio station at Leafield, Oxfordshire, is to become fully operational in the summer of this year.

This station, costing over £1M, has been designed to handle the multi-destination press news service and the expanding public radio telegraph and telex services.

It contains six 85kW transmitters for news services and twelve 30kW transmitters for the radio telegraph and telex services.

It replaces the original Leafield station built in 1921.

The Industrial Control and Electronics Division of the British Electrical and Allied Manufacturers' Association (BEAMA) has set up a Committee to discuss the overall concept of applying control systems to ships and marine installations and to prepare in conjunction with other associations, internationally acceptable data in respect of operating parameters, component life and other essential factors affecting the operation of the control equipment at sea.

The Committee, which consists of marine specialists from BEAMA and the British Shipbuilding Research Association, is currently engaged in an analysis of the recommendations for automation in ships recently issued by Lloyd's Register of Shipping, Bureau Veritas and Norske Veritas. Following discussions with the Chamber of Shipping of the United Kingdom, the Committee is keeping in touch with the views of United Kingdom shipowners through the appropriate British Ship Research Association Committees and is also in contact with other classification Societies and marine organizations throughout the world.

BEAMA has already produced a guide to the specification and purchase of electronic equipment for industrial systems and this is now being brought up to date to include specific references to marine equipment in the closest collaboration with the British Shipbuilding Research Associations. Copies of this booklet are available from the Information Division, BEAMA, 6 Leicester Street, Leicester Square, London, W.C.2, at £1 1s. 0d. per copy.

The Aerospace and Defence Group of G.E.C. (Electronics) Ltd has received a development contract from the G.P.O. for modulator and demodulator equipment for use with Goonhilly No. 2 aerial and future development at the Satellite Communications Station in Cornwall.

The equipment will be a key part of a major installation at the new No. 2 aerial at Goonhilly Down, near Lands End, Cornwall, which is due for completion late in 1967.

Special purpose wide deviation modulator equipment allied to a low threshold feedback demodulator will raise the standards of clarity on trans-Atlantic audio and video links connecting directly with the British telephone communications network with benefits to both commercial users and television and radio audiences.

The equipment will take a group of voice or a single video signal from a microwave link and convert it to a modulated r.f. signal to drive the transmitter. Incoming weak r.f. signals from a satellite will be converted into a clear voice or video signal on the telephone network.

The Integrated Electronic Systems Division of Standard Telephones and

Cables Ltd has supplied Train Describers and Signal Post Telephone Systems to British Railways for their signalling systems on the new Euston to Manchester and Liverpool electrified lines.

The Train Describers, located in key signal boxes show the disposition of trains within the area controlled by the signal box. Each train's individual four digit code is displayed in the appropriate signal berth display on a track diagram. As a train moves through the area its code is automatically stepped from one berth to the next in accordance with its progress, enabling the signalmen to see at a glance where any given train is at any given time.

Signal post telephones, mounted in special weatherproof housings on or near specific signal posts, enable the driver of a train brought to a halt by a signal to communicate directly with the appropriate signalman. Each signal post telephone on a given circuit is allocated a separate identification code which is displayed on a panel in the signalbox on receipt of a call. A safety feature is incorporated whereby all other signal post telephones on the circuit are inoperative from the time a telephone call is commenced, until it is re-set by the signalman. This prevents drivers acting on instructions not intended for themselves.

The 128th Annual Meeting of the British Association for the Advancement of Science is to be held at the University of Nottingham on 31 August to 7 September this year.

Full details regarding programme and registration are obtainable from the Association at 3 Sanctuary Buildings, 20 Great Smith Street, London, S.W.1.

The Marconi Company Limited is to supply Aalborg, the second largest airport in Denmark with a 50cm radar display system together with a number of new devices designed to eliminate clutter and interference.

A high-power version of the same radar system was supplied in 1960 to Kastrup, the principal airport in Denmark.

The installation at Aalborg will contain a daylight radar display system, which will enable the Approach Radar Controllers to work in conditions of bright daylight. Main and standby 5in daylight displays will also be installed in the 'glass-house' on top of the control tower, at both Aalborg and Kastrup.

An important feature of the new system, is that it will operate on almost the same radio frequency as the existing radar at Kastrup. This has been done to conserve space in the radio frequency band, and to minimize the effects of interference from stations on adjacent frequencies. It has been possible by the incorporation of pulse repetition frequency (p.r.f.) discriminators at both Aalborg and Kastrup. The two radar sites will use different pulse repetition frequencies, and each receiving system will be designed to receive signals at its own p.r.f. only.

The Aalborg installation will also include a pulse length discriminator, which will further improve the quality of the radar pictures, by rejecting all responses in the receiver which are of a greater length than those coming from an aircraft. This will further reduce random interference and clutter on the displayed pictures.

The British Standards Institution has published a new British Standard No. B.S.3968 "T-Track Magnetic Tape for Data Interchange recorded at 200 rows per inch" dealing with standard magnetic tapes for computers and data-processing.

Although British Standards have been published in the past covering magnetic recording tape and spools, these have dealt with several different widths of tape. The new standard, however, is concerned entirely with recorded tape, ½in wide, having only 7 tracks and a density of recording of 200 rows per inch across these tracks and intended for general national and international interchange. Spools for use with this tape are also covered by the standard.

The draft of a companion specification for 9-track tape recorded at 800 rows per inch is now being commented on by industry. Both standards are in accord with the forthcoming ISO (International Organization for Standardization) Recommendations on the subject; the recommendation for 7 track tape is being based on the new British Standard.

Copies of the B.S.3968, may be obtained from BSI Sales Branch, 2 Park Street, London, W.1. Price 10s. each (Postage 1s. will be charged extra to non subscribers).

The Third Congress of the International Federation of Automatic Control (IFAC), organized by the United Kingdom Automation Council with the co-operation of the Secretariat of IFAC and its National Member Organizations is to be held in London on 20 to 25 June this year.

More than 2000 delegates from 35 countries are expected to attend and the subjects will range from space technology to the central control of the Montreal underground railway system.

Full details regarding programme and registration from the IFAC Congress Secretariat c/o The Institution of Electrical Engineers, Savoy Place, London, W.C.2 or from The National Member Organizations.

Fellowship of the Royal Society has been awarded to the following:

Dr. K. G. Budden, Cavendish Laboratory, University of Cambridge for his contributions to the theory of the propagation of electromagnetic waves in a non-uniform anisotropic medium, particularly of long radio waves in the ionosphere.

D. G. King-Hele, Royal Aircraft Establishment Farnborough for his mathematical work on near-earth satellite orbits, and for the inference of geophysical data from its comparison with observations.

Prof. J. D. McGee, Imperial College of Science and Technology, University of London for the application of physics to developing electron-optical instrumentation, television camera tubes, infra-red image convertors and image intensifiers.

A British Combined Stand, sponsored and organized by the London Chamber of Commerce, is to be made available at the Electronica Exhibition, Munich on 20 to 26 October this year.

The stand will have the usual Board of Trade Joint Venture facilities and will be available to all British Companies associated with the electronic industry.

Further information is obtainable from the U.K. Agents for the Organizers, Exhibition Consultants Ltd, 11 Manchester Square, London, W.1.

"SIMA: its role; its structure; its activities" is the title of the revised handbook issued by the Scientific Instrument Manufacturers' Association of Great Britain.

As well as a complete list of members, the committees and specialized groups representative of the industry are listed in the handbook, together with their terms of reference, with the names of members and their respective companies. Specialized groups deal with the needs of instrumentation in: agriculture, education, medicine, nucleonics, space research, traffic engineering, ultrasonics, the textile industry, machine tools and production equipment. Another special group covers the activities of SIMA members in northern England to ensure close liaison between that area and the administration in London. This arrangement has the particular advantage that the Association as a whole is aware of any special problems peculiar to the north.

The AEI Telecommunications Circuit Development Laboratory has produced a four channel speaking clock capable of producing simultaneous time announcements in up to four languages at 10, 15 or 20-second intervals, accurate to within half a second a month. Every channel, one per language, can feed up to 60 speech paths.

The speaking clock equipment is contained in a pressurized cabinet to ensure a dust-free atmosphere, and is suitable for operation in all climates.

A clock of this type is being supplied to Dar Es Salaam to give time announcements in English and Swahili.

The Telefunken Company of Berlin is to supply the Cracow Airport, the second international airport in Poland with a stationary Precision Approach Radar PAR-T 3 of the latest design.

The Company has already installed a similar equipment at Warsaw Airport, which in conjunction with a Telefunken airport surveillance radar constitutes a complete GCA system.

The Solartron Electronic Group Ltd of Farnborough, Hants, has received an order valued at £40 000 from the Australian Department of Civil Aviation for the supply of a 24-target air traffic control simulator capable of simulating supersonic speeds.

The simulator has a 'playing area' 400 miles square with an altitude of 50 000ft and air speeds of up to 3 000 knots for each aircraft.

Apart from the training roles, the prime purpose of the simulator will be the evaluation of a.t.c. procedures, in particular the approach procedures on major airfields such as Tullamarine (Melbourne), Mascot (Sydney), Brisbane, Adelaide and Perth.

The Scientific Instrument Manufacturers' Association, in co-operation with the Ministry of Health, the Board of Trade and the British Hospitals Export Council, is to hold a British Hospital and Medical Equipment Exhibition at the Phoenicia Hotel, Beirut, Lebanon on 11 to 17 May this year.

The exhibition has been timed to coincide with the Arab world's 1966 medical assembly with the object of expanding the market for British medical equipment in that area.

Of the fifty companies taking part nearly a quarter are members of SIMA. Between them they will show a wide range of British medical instrumentation, including blood cell counters, centrifuges, model linear accelerators, autoclaves, laboratory incubators, chromatography apparatus and radiotherapy equipment.

A.E.I. Ltd, G.E.C. Ltd and the Plessey Co. Ltd have formed a consortium to build and sell complete communications satellite ground terminals. The market for such terminals for civil use is estimated to be £100M and the consortium has been formed to capture a major share of this business for Britain.

Joint engineering teams are already at work on designs to meet the requirements of this international market.

A.E.I., G.E.C. and Plessey together have unrivalled engineering and commercial experience in the design, construction and sale of catellite communications ground terminals throughout the world including large aerial structures, microwave engineering and telecommunications.

The consortium feels it will thus be able to meet the rapidly rising world demand for satellite terminals.

The Air Traffic Management Group of General Precision Systems Ltd is to develop a fast-time simulator model for Eurocontrol, the European organization for the safety of air navigation.

The model will be 'general purpose' in nature, and capable of representing a large volume of air space containing complex route structures serving at least two major air traffic generating sources, together with the associated air traffic control systems environment. Under the contract, the model will be used in an investigation into the advantage to be derived from the use of area coverage (as opposed to point source) navigation systems within the air spaces, which come under the jurisdiction of Eurocontrol, i.e. West Germany, Belgium, France, Holland, Ireland, Luxembourg and the United Kingdom.

In the autumn of 1967 this model, in the form of a set of computer programmes, will be taken over by Eurocontrol for loading into their own computers at the Eurocontrol Experimental Centre at Bretigny, near Paris. Here, the model will be used for long-term future air traffic control system design studies.

Eurocontrol is believed to be the first major air traffic control organization to make extensive use of the new and sophisticated techniques in fast-time simulation, which can produce an advance assessment of overall air traffic control system performance under a wide range of conditions.

Standard Telephones and Cables Ltd has received an order valued at £45 000 from the Independent Television Authority for the supply of Band III television translators.

The order covers FTU 5-A translator units and the associated power and control equipment. The units are fully solid state and are suitable for the British monochrome 405-line system. The output power is 2W. STC also manufactures a similar translator with a power output of 1W for the 625-line system for monochrome or colour transmission.

The ITA contract follows closely upon a BBC order placed with STC for a number of Band IV and V translators which will serve the Midlands area.

The Aircraft Equipment Department of Ferranti Ltd, Bracknell, Berkshire, has developed a new type of artificial horizon with virtually unlimited pitch range and providing an uncluttered display of attitude.

This is to meet the requirement for high flying, high performance aircraft.

Known as the type FH14, it contains its own vertical gyro and features a new electrically driven spherical attitude display, the first of its type developed by a British manufacturer. This provides for a linear display of pitch attitude between $\pm 85°$ and full freedom in roll.

The display is in two colours, blue and black, and on the ground (black) area, a perspective pattern gives depth to the display. In addition, a pattern of pitch lines increasing in length from the horizon line provide a self-evident indication of horizon direction when this is not in view and gives additional roll guidance at high pitch angles. The instrument, which is designed to meet full A.R.B. requirements, is fitted in a standard size 3 A.T.I. case.

Morganite Resistors Ltd of Jarrow, County Durham, has signed a licence agreement with Beckman Instruments Inc, of California, to manufacture trimming potentiometers and other devices embodying the Beckman CERMET (thick film) resistive glazes.

The construction of the CERMET potentiometer involves the screen printing and subsequent firing of a glass/noble-metal composition on to a ceramic substrate making a potentiometer immune to failure through atmospheric attack.

A high power rating is also ensured.

The Marconi Co. Ltd has been appointed exclusive distributor in the United Kingdom for the range of language laboratories manufactured by Cédamel of Paris.

The complete range of equipment is being marketed by a newly formed Educational Systems Group within the Closed Circuit Television Division of The Marconi Company, with headquarters at Basildon, Essex.

A number of different types of language laboratory is available, including two new transistorized systems, the Educomatic B and the Rexmatic. Both systems provide facilities for classes of up to 48 students with extreme simplicity of operation and high reliability. In addition the Rexmatic system provides for full remote control of the students' booths from the teachers' console.

The Guided Weapons Division of the British Aircraft Corporation, Bristol, has received a contract from the Ministry of Aviation (Navy Department) to provide support to the Chief Polaris Executive, Navy Department for the commissioning of certain aspects of the U.K. Polaris Weapon System.

The Division will be responsible for testing and tuning the Polaris missile system, i.e. the missile test and readiness, the guidance system and related equipment as installed in the submarine.

It will also be responsible for analysis of tests of the complete weapon system before hand-over to the Royal Navy, and will provide a consultancy service to the Chief Polaris Executive in respect of such weapon system engineering problems as may arise.

NOUVELLES Réalisations

Traduction des pages 330 à 335

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

DÉTECTEUR SENSIBLE AUX PHASES

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustration à la page 330)

La société Brookdeal Electronics Ltd produit un nouveau détecteur transistorisé sensible aux phases couvrant la gamme de fréquences de 10 Hz à 300 kHz. Ce détecteur comporte son propre appareil de lecture et fournit des sorties permettant d'actionner une gamme étendue d'enregistreurs à plume, de voltmètres numériques etc.

Le degré élevé de linéarité et la faible dérive nulle permettent d'utiliser cet instrument à des niveaux de signaux très faibles en présence de tensions de bruit beaucoup plus élevées. Il est particulièrement utile comme détecteur à tension alternative dans les systèmes de précision. Une attention spéciale a été accordée à la réduction au minimum des opérations de réglage; on a ainsi éliminé le choix de la gamme de fréquence et le réglage précis de la tension de référence.

Le circuit de contrôle qui comprend un instrument de mesure à grand cadran avec zéro central ou à gauche comporte des dispositifs comprenant des constantes de temps de commutation allant de 0,03 à 3 sec, des atténuateurs à bouton-poussoir, un filtre facultatif pour le fonctionnement à basse fréquence et des commandes de suppression de zéro.

L'appareil peut être monté sur châssis de l'administration des postes de 48,2 cm et il peut être fourni dans un coffret à instruments ou dans un coffret à deux instruments avec préamplificateur approprié.

EE 93 751 *pour plus amples renseignements*

RELAIS

Painton & Co. Ltd, Kingsthorpe, Northampton

(Illustration à la page 330)

La société Painton & Co. Ltd vient de conclure un accord de licence exclusif avec la société Chauvin Arnoux de Paris pour la fabrication de la totalité de sa gamme de relais OK en Grande Bretagne. La société Painton reçoit les droits exclusifs de vente et de distribution de ces relais non seulement pour la Grande Bretagne mais également pour l'Irlande, l'Australie, la Nouvelle Zélande, le Canada et l'Afrique du Sud.

Ces relais, dont la puissance nominale maxima de courant est de 10 A par contact, avec quatre contacts à permutation, et dont la durée de vie mécanique est de plusieurs millions d'operations, couvrent une gamme étendue d'applications. Des modèles peuvent ainsi être obtenus pour la rupture de circuit, le retard de temps, les applications logiques et diverses autres utilisations plus spécialisées.

Grâce à l'emploi de nouvelles techniques de réalisation, le relais est maintenu en réglage permanent et le circuit magnétique aussi bien que le circuir mécanique sont extrêmement sûrs. Toutes les variantes de ces relais sont incorporées à un coffret transparent à l'épreuve de la poussière et de dimensions courantes. Il existe des versions pour courant alternatif et pour courant continu. Une gamme complète d'accessoires de montage est prévue et les connexions de base sont du type en forme de bêche convenant pour le montage sur douille ou pour les connexions du type à poussoir.

EE 93 752 *pour plus amples renseignements*

VOLTMÈTRES NUMÉRIQUES

Distributeurs: Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustration à la page 330)

Un nouveau voltmètre numérique d'une précision de 0,1 %, de construction modulaire et de style ergonomique, a été réalisé par la société Kynmore Engineering Co. Ltd. Portant le nom de voltmètre numérique Vénus, cet appareil est construit par la société française Schneider et il est prévu en deux modèles standard comportant tous deux un choix de raffinements qui, ajoutés à la gamme d'accessoires, les rendent propres à la mesure de haute qualité.

L'un des modèles standard, le Vénus 123, comporte trois gammes de mesure: $-9{,}99$ à $+9{,}99$ V, $-99{,}9$ à $+99{,}9$ V et -999 à $+999$ V c.c. Ce modèle fournit une indication automatique de la décimale et de la polarité et sa précision est de 5×10^{-4} de la lecture sur la totalité de la gamme ± 1 chiffre.

L'autre modèle standard, le Vénus 124, comprend un seul chiffre supplémentaire donnant une indication des surcharges. L'indication maxima dans chacune des trois gammes est donc de 10,00, 100,0 et 1 000 V respectivement, mais des valeurs supérieures sont indiquées.

La surcharge admissible est d'environ 25 % pour les deux modèles, à l'exception de la gamme de 1 000 V. L'impédance d'entrée minima est de 20 MΩ avec 1 000 MΩ facultatifs sur la gamme de 10 V. La température de fonctionnement nominal est de $+5$ à $+45°$ C. La mesure est automatiquement renouvelée toutes les secondes.

Les accessoires facultatifs comprennent une sortie d'imprimeur, et les deux modèles peuvent fonctionner sur imprimeur Schneider (également fourni par Kynmore) ou avec un instrument Schneider auxiliaire (le "Mars") qui le convertit en un multimètre numérique. L'appareil Mars fournit deux gammes de tension continue inférieures allant jusqu'à 100 mV sur la totalité de l'échelle, quatre gammes de tension alternative de 300 mV à 300 V efficaces sur la totalité de l'échelle, ainsi que des gammes de résistance. La précision, la sensibilité et l'impédance d'entrée dépendent de la gamme utilisée mais ces caractéristiques sont comparables à celles du voltmètre Vénus.

Le principe de fonctionnement est la modulation de la durée d'impulsions, le voltmètre Venus étant muni d'un détecteur différentiel à modulations par impulsions contrôlé par un oscillateur piloté au quartz. L'indication automatique de polarité et de tension est effectuée par logique interne, et l'indication apparaît sur des tubes à cathodes froides. Le choix de la gamme ou de l'étalonnage

ELECTRONIC ENGINEERING occupera le Stand No N 816 à l'Olympia de Londres durant le Salon des Instruments, de l'Électronique et de l'Automation du 23 à 28 mai, et les visiteurs y seront les bienvenus.

se fait au moyen de clavettes de "piano" mais la référence est réglée intérieurement. Une fois réglée, la référence demeure stable pendant de très longues périodes suivant la période de chauffe initiale de dix minutes.

EE 93 753 *pour plus amples renseignements*

CONDENSATEURS D'APPOINT À DISQUE EN CÉRAMIQUE

Steatite Insulations Ltd, Hagley House, Edgbaston, Birmingham, 16

(Illustration à la page 331)

La société Steatite Insulations Ltd. vient d'ajouter quelques nouveaux modèles à sa gamme de condensateurs d'appoint à disque en céramique subminiature.

Notre gravure montre sur le côté gauche le condensateur type S-Triko 03 conçu pour répondre aux besoins précis d'utilisateurs militaires et professionnels. Ce condensateur est conforme à la spécification MIL-C 81A est il est livrable dans les gammes de capacitance suivantes:

P100	2	à	3·5	pF
N033	2·5	à	6	pF
N075	3	à	10	pF
N470	3·5	à	13	pF
N750	4·5	à	20	pF
N1500	7	à	35	pF

Sur le côté droit, on voit le condensateur à disque en céramique type 5S-Triko 03 dont il existe une série de modèles dans les gammes de capacité suivantes:

P100	1·5	à	2·5	pF
N033	2·5	à	6	pF
N075	3	à	8	pF
N470	3·5	à	11	pF
N750	4·5	à	15	pF
N1500	6	à	22	pF

EE 93 754 *pour plus amples renseignements*

CHRONOMÈTRE NUMÉRIQUE

Darang Electronics Ltd, Restmor Way, Hackbridge Road, Hackbridge, Surrey

(Illustration à la page 331)

Le chronomètre type 661D fait partie de la nouvelle gamme de chronomètres numériques électroniques "Digicron" de la société Darang Electronics Ltd. La conception est basée sur un système de plaquettes à fiches à bas prix et sur l'emploi de méthodes extrêmement sûres à cathodes froides. De nombreuses variantes sont possibles pour répondre aux nécessités particulières des utilisateurs.

Un affichage en ligne à cinq chiffres est assuré pas les tubes d'affichage numériques à gaz. La période de chronométrage couverte est de 59 min 59,9 sec en unités de 0,1 sec. Sur le modèle standard, la fréquence d'alimentation secteur est utilisée comme source de chronométrage, mais on peut également utiliser une source de chronométrage extérieure. On peut ajouter, moyennant une supplément de prix, une source de chronométrage à cristal dans un four à température contrôlée.

L'arrêt et la mise en marche s'effectuent par bouton-poussoir sur le panneau frontal ou par contacts extérieurs à distance. La longueur de câble allant aux contacts à distance peut dépasser 1600 m.

Le réenclenchement est automatique par la mise en action des contacts de "marche". En variante, on peut ajouter un certain nombre de périodes de temps et afficher le temps total.

L'alimentation requise est de 95 à 130 V ou de 190 à 260 V, 50 Hz.

EE 93 755 *pour plus amples renseignements*

INSTRUMENT DE MESURE GALVANIQUE DE LA RÉSISTANCE DE LA PEAU

Short Brothers & Harland Ltd, Precision Engineering Division, Castlereagh, Belfast, N. Ireland

(Illustration à la page 331)

L'instrument de mesure de la résistance de la peau Short est un appareil autonome fonctionnant sur courant alternatif de secteur de 230 V. Il permet à l'opérateur de mesurer:

(a) La résistance de la peau d'un sujet
(b) Les changements de résistance qui se produisent par suite de divers facteurs, soit neutres soit provoquant un état d'anxiété.

Les conducteurs d'électrodes de peau sont branchés sur deux prises sur le panneau frontal et un courant constant de 10 μA traverse la partie de la peau soumise à l'essai. L'instrument de mesure du panneau frontal indique la résistance de base et il n'est pas affecté par les variations de résistance. Trois gammes sont prévues, qui donnent des lectures sur la totalité de l'échelle de 100 kΩ, 200 kΩ et 500 kΩ. Un jack du panneau frontal est utilisé pour relier un enregistreur à plume qui n'enregistre que les changements de résistance et indique zéro lorsque aucun changement n'a lieu.

Cette séparation entre les changements de résistance de base et de résistance ordinaire est un gros avantage par rapport à de nombreux autres instruments de mesure galvaniques de résistance de la peau qui exigent de la part de l'opérateur un réglage des potentiomètres pour l'analyse de la résistance de base de manière à pouvoir observer les changements. Le seul réglage nécessaire avec l'instrument Short est un réglage quotidien des commandes de zéro de l'amplificateur sur le panneau frontal.

EE 93 756 *pour plus amples renseignements*

GÉNÉRATEURS D'IMPULSIONS

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 331)

Deux nouvelles versions du générateur d'impulsions Datapulse 110 constitué de corps solides sont maintenant construites au Royaume-Uni par la société Dynamco Instruments Ltd. Il s'agit du modèle 110A qui comporte un temps de montée supérieur et la possibilité de déclenchement externe (sans augmentation de prix), et le modèle 110P qui est une version à programme du 110A.

Le 110A offre un temps de montée minimum de 4,5 nsec (4 nsec typiques) et le déclenchement extérieur à partir de 1 V. Le taux de répétition varie de 4 Hz à 40 MHz. Les temps de transition linéaire sont à variations séparées. Des sorties simultanées positives ou négatives sont prévues jusqu'à 10 V dans 50 Ω avec une atténuation allant jusqu'à 70 dB, soit dans le mode à impulsions simples soit dans le mode à impulsions doubles. La durée des impulsions varie de 10 msec à 5 msec, le retard d'impulsion varie de 10 nsec d'avance à 50 msec de retard. Le cycle de régime caractéristique est de 85%. Une commande variable de la ligne de base de tension continue permet d'obtenir des impulsions de sortie complémentaire à n'importe quel niveau entre ± 10 V. Des commandes de gammes étendues permettent le contrôle de tolérance, facilitent la répétition des résultats d'essais et permettent la simulation exacte des formes d'ondes triangulaires et de rampe pour la mesure d'inductance d'impulsions et les essais d'amplification. La sortie d'impulsions simples permet la localisation de défauts dans les circuits logiques. Les étages de sortie ne risquent pas d'être endommagés par une combinaison quelconque de réglages de commandes du panneau, de court-circuits de n'importe quelle durée ou de tensions de retour allant jusqu'à ± 10 V. La capacité nominale de fonctionnement de l'appareil couvre des températures ambiantes de 0° à 50° C.

Les générateurs d'impulsions à programme de la série 110P permettent la programmation automatique et/(ou) contrôlée à distance de tous les paramètres d'impulsions, y compris les temps de montée et la compensation de tension continue de la ligne de base. Des essais de composants à action rapide, tel que l'essai de circuits intégrés, peuvent être effectués par l'emploi d'accessoires tels que les lecteurs de cartes, les bandes de papier perforé ou la programmation par disque magnétique.

EE 93 757 *pour plus amples renseignements*

VOLTMÈTRE D'ANGLE DE PHASE

Smiths Industries Ltd, Aviation Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustration à la page 331)

La division d'aviation de la société Smiths Industries Ltd produit maintenant un nouveau voltmètre compact d'angle de phase de 400 Hz constitué de corps solides et comprenant de nombreux perfectionnements utiles. Cet instrument mesure 21,59 cm de long × 15,24 cm de haut × 20,32 cm de profondeur et sa précision est de ±3% d'amplitude et de ±1 degré en phase.

Il peut être utilisé pour présenter soit les composants résolus soit la phase et le module d'entrée; la sensibilité prévue dans 12 gammes varie entre 300 V et 1 mV de déviation sur la totalité de l'échelle. Le signal de référence peut être appliqué à n'importe quelle valeur entre 300 V et 100 mV, ce qui rend inutile un réglage manuel du niveau. La commande à réglage unique ne doit être ajustée que lorsqu'il se produit un changement de la fréquence de signal. Les entrées de signal et de référence sont accouplées à un transformateur, ce qui donne le rapport élevé de rejet de mode commun de 10^5:1 exigé pour des mesures différentielles et de pont valables. Des filtres sont inclus dans les canaux de référence et de signaux pour fournir le degré nécessaire de rejet d'harmoniques.

Un facteur de surcharge de 10 : 1 a été prévu de sorte que des composants mineurs peuvent être mesurés en présence d'un grand signal total. Une lampe d'avertissement s'éclaire lorsque ce facteur est dépassé.

L'instrument fonctionne sur n'importe quelle alimentation entre 100 V et 250 V, 50 à 400 Hz. Nul réglage n'est nécessaire pour les changements de tension ou de fréquence, ce qui réduit le risque d'erreurs lorsque différents types d'alimentation sont employés.

EE 93 758 *pour plus amples renseignements*

ÉMETTEUR-RÉCEPTEUR TRANSISTORISÉ

G.E.C. (Electronics) Ltd. Communications Group, Spon Street, Coventry

(Illustration à la page 332)

L'émetteur-récepteur RC/420/TR qui ne pèse que 7 kg environ fournit des sons vocaux à grande distance ou des communications par ondes entretenues dans la bande HF de 2 à 18 MHz. Il peut être utilisé comme station fixe ou mobile et se loge facilement dans un tableau de bord de voiture.

L'émetteur-récepteur RC/420/TR comporte un certain nombre de perfectionnements des plus intéressants. La plupart des circuits se trouvent autour d'un seul type de transistor mésa au silicium. Des transistors au germanium ne sont utilisés que dans le bloc d'alimentation et les étages acoustiques. On améliore ainsi la fiabilité et la facilité d'entretien. Quatre canaux pilotés par quartz dans la bande de 2 à 18 MHz sont prévus pour le fonctionnement sur bande latérale unique ou sur bande latérale double. Un filtre facultatif de bande latérale inférieure peut être monté pour fournir huit canaux de fréquence de fonctionnement. Un circuit clarificateur sous le contrôle de l'opérateur minimise la nécessité d'utiliser des fours à quartz à haute consommation de courant, réduisant ainsi le drainage de batterie. Toutes les plaquettes de circuit et les pièces de clavettes sont traitées contre la pousse fongueuse.

L'appareil est fourni complet avec un microphone dynamique. L'assortiment d'accessoires comprend un dispositif pour la transmission directe aux lignes téléphoniques; un bloc d'alimentation en alternatif de 110/220 V; un accordeur d'antenne pour les fréquences multiples, un bâti de montage et une installation d'antenne pour les applications mobiles. Un appareil téléphonique de table classique ainsi qu'un équipement téléimprimeur peuvent être fournis.

EE 93 759 *pour plus amples renseignements*

MILLIVOLTMÈTRE DE DISTORSION ET DE DÉCIBELS

Laboratoire Electro-Acoustique, 5 Rue Jules Parent, Rueil (S & O), France

(Illustration à la page 332)

Le "Qualiscope" est un instrument de précision de type universel réunissant en un seul élément les moyens de mesurer les principales propriétés d'un circuit. Les paramètres mesurés sont: la distorsion non-linéaire (pourcentage); la tension (mV); le niveau de tension (dB) et le niveau de bruit.

Le distorsiomètre couvre des fréquences fondamentales de 10 Hz à 100 kHz et des fréquences d'harmoniques de 10 Hz à 250 kHz, la sensibilité maxima étant de 0,1% de déviation sur la totalité de l'échelle.

Le voltmètre-décibelmètre couvre la gamme de fréquences de 2 Hz à 400 kHz, la réponse de fréquence étant à ±0,2 dB près de 20 Hz à 100 kHz et à ±2 dB près de 2 Hz à 400 kHz. La gamme de tension s'étend de 1 mV à 300 V de déviation totale. Un préamplificateur supplémentaire (gain +20 dB) porte la sensibilité à 100 µV de déviation totale.

L'instrument comprend un tube cathodique incorporé pour l'affichage des chiffres de Lissajous.

EE 93 760 *pour plus amples renseignements*

COMPTEUR D'IMPULSIONS AUTOMATIQUE

L'Electronique Appliquée, 96-100 Rue Maurice Arnoux, Montrouge (Seine), France

(Illustration à la page 332)

Ce compteur transistorisé comporte un dispositif d'affichage de six tubes "Nixie" et un régistre électromécanique à six chiffres avec réenclenchement par bouton-poussoir. Le mode de fonctionnement voulu est choisi par bouton-poussoir sur le panneau frontal.

Le compteur peut compter des impulsions à l'intérieur d'un seuil réglable de 0,1 V à 10 V, la résolution d'entrée étant de 1 µsec. L'impédance d'entrée pour les impulsions positives est de 10 kΩ et pour les impulsions négatives de 2,5 kΩ.

Pour la mesure de fréquence, le temps de comptage est d'une seconde et le temps de lecture les deux secondes suivantes.

Le compteur peut être arrêté automatiquement après comptage présélectionné de n'importe quel chiffre jusqu'à n 10 p n étant réglable de 1 à 9 et p de 0 à 6. En variante, le compteur peut être arrêté automatiquement après des durées présélectionnées de 1 à 10^6 sec. Si la durée et le comptage sont tous deux présélectionnés le compteur s'arrêtera à celui des deux qui est atteint le premier.

L'étalon de fréquence est assuré par un cristal de quartz de 100 kHz.

Une sortie est prévue pour actionner un imprimeur.

EE 93 761 *pour plus amples renseignements*

SERVOMOTEUR

Servomex Controls Ltd. Crowborough, Sussex

(Illustration à la page 333)

Ce nouveau servomoteur biphasé de 50 Hz de petites dimensions avec générateur tachymétrique et boîte de vitesses démultiplicatrice est de construction simple et d'un prix remarquablement réduit. Il a été conçu pour remplacer le type de servomoteur classique essentiellement prévu pour des fins militaires dont la performance est remarquable mais dont le prix est également très élevé.

Le matériel employé pour le nouveau moteur Servomex, ainsi que sont exécution, sont de la plus haute qualité et il peut donc être incorporé à des mécanismes d'asservissement d'instruments de la meilleure qualité.

Le moteur est du type quadripolaire à barres de rotor hélicoïdales et à générateur tachymétrique. Il existe deux modèles, le SM 95 et le SM 135, à enroulements et à paliers différents mais dont le rendement est sensiblement le même. L'arbre de sortie du modèle SM 95 s'engage dans un palier en bronze poreux imprégné de lubrifiant au silicone, tandis que le modèle SM 135 a un arbre de sortie s'engageant dans un roulement à billes de précision et comprenant un ergot usiné avec précision.

Les bobines sont imprégnées au vide selon la spécification RCS 214. La boîte de vitesses dont le rapport de démultiplication est de 40 : 1 utilise des engrenages en acier et au nylon et comprend un embrayage qui patine. Cet embrayage vise à empêcher tout endommagement ou tout impact.

L'arbre de sortie a une vitesse maxima sans charge de 60 tours/min, le couple de décrochage étant de 850,5 gr/2,5 cm. Les deux modèles sont en alliage d'aluminium et peuvent être utilisés dans une gamme de température de −10° à +70° C.

Le SM 95 peut être entraîné par un amplificateur à changement de phase de 90°, les bobinages de référence du moteur et d'excitation du générateur étant alimentés sur phase nulle de 6,3 V c.a. Le bobinage de référence du moteur du SM 135 a été conçu pour être alimenté sur 230 V efficaces à travers une capacitance de 1 µF à 1,4 µF. Ainsi, le courant dans le bobinage entraîne la tension d'alimentation à 90°, permettant d'amorcer les bobinages de commande à partir d'un amplificateur de déphasage nul.

Le servomoteur mesure 7,62 cm × 11,43 cm × 6,03 cm et son poids est d'une vingtaine de kilos.

EE 93 762 *pour plus amples renseignements*

CONDENSATEURS À DISQUE EN CÉRAMIQUE

Centralab Ltd. Northway House, Great North Road, London, N.20

Il s'agit ici d'une nouvelle gamme de condensateurs à disque en céramique à bas prix dont le format réduit et la performance constituent des facteurs appréciables. Les trois valeurs de capacitance sont: 0,01, 0,02 et 0,05 microfarads pour des diamètres de disque de 0,297, 0,359 et 0,515 pouces respectivement, et une épaisseur ne dépassant pas 0,125 pouce.

La tension de régime de la gamme est de 50 V c.c. Tous les formats ont un espacement de conducteur de 0,25 pouce sauf le plus grand dont l'espacement est de 0,375 pouce, les conducteurs étant fournis soit droits soit avec une boucle à l'extrémité et coupés pour l'emploi dans les plaquettes imprimées. Tous les conducteurs sont revêtus de soudure pour faciliter leur assemblage.

Chacun des éléments de la gamme est fourni avec une valeur de tolérance de +80%, −20% ou ±20%.

EE 93 763 *pour plus amples renseignements*

CONTRÔLEUR DE TEMPÉRATURE

Société D'Etudes et Recherches Electroniques, 15 Rue Villebois-Mareuil, 92-Colombes (Seine), France

(Illustration à la page 333)

Le nouveau contrôleur de température SERELEC assure le contrôle proportionnel, dérivatif et intégral ainsi que la régulation rigoureuse—typiquement de 0,2° C/500° C—malgré les fluctuations de tension secteur, de température ambiante ou tout autre dérangement extérieur.

Il comprend trois étages amplificateurs magnétiques, ainsi que deux thyristors pour le contrôle de la sortie.

EE 93 764 *pour plus amples renseignements*

FILTRES YIG ACCORDABLES ÉLECTRONIQUEMENT

Microwave Associates International Inc.; European Office: Cradock Road, Luton, Bedfordshire

(Illustration à la page 333)

La société Microwave Associates a introduit une nouvelle série de filtres YIG accordables électroniquement dans la gamme de fréquences de 125 à 12 000 MHz. Ces filtres ont un affaiblissement d'insertion très réduit, un bon rejet hors-bande et ils sont exempts de réponse non-sélective.

Pesant moins d'un kilo, leur rapport d'amplitude de tension maxima est de 1,5 et ils répondent à toutes les conditions de la spécification MIL-E-5400, Class II.

Ces filtres sont constitués par des sphères cristallines polies en yttrium, telles qu'utilisées dans les éléments d'accord gyromagnétique.

Les amplificateurs d'entraînement actionnent, au moyen d'un signal de 0 à 10 V émanant d'une source à haute impédance, les filtres de l'extrémité inférieure à l'extrémité supérieure de leur gamme respective. Des blocs d'alimentation et de contrôle de température peuvent également être fournis si nécessaire.

Ces éléments ont été conçus pour assurer la présélection et la post-sélection pour n'importe quel type de récepteur exigeant la sélection par analyse rapide ou par signal automatique, ou encore pour le traitement des signaux.

EE 93 765 *pour plus amples renseignements*

TRIMMERS

Steatite Insulations Ltd. Hagley House, Hagley Road, Edgbaston, Birmingham, 16

(Illustration à la page 333)

Le trimmer à vis type 1250b vient d'être ajouté à la gamme des condensateurs shunt d'équilibrage de la société Steatite Insulations Ltd. Conçu pour montage sur circuit imprimé, ce trimmer a une capacité maxima de 10 ou de 20 pF; la capacité minima dans les deux cas étant de 2 pF. La loi de capacité est linéaire dans un angle d'accord de 6 × 360°. Le facteur de puissance à 1 MHz est supérieur à 0,002 et la tension de régime est de 250 V c.c. Les dimensions hors-tout du trimmer sont de 15,5 mm × 7,5 mm.

EE 93 766 *pour plus amples renseignements*

AMPLIFICATEUR DE TENSION CONTINUE

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Illustration à la page 333)

La société S.E. Laboratories (Engineering) Ltd a réalisé un amplificateur à tension continue à entrée différentielle entièrement flottante, à gain élevé, large bande et faible bruit. Cet oscillateur, type S.E. 4200, peut être utilisé avec des enregistreurs de données magnétiques et oscillographiques, des systèmes modernes d'acquisition de données, etc.

La conception du S.E. 4200 lui permet d'amplifier des signaux de niveau très bas, tels que les signaux provenant de thermocouples, d'extensomètres et de dispositifs analogues.

Le nouvel amplificateur se caractérise par une haute stabilité qu'on n'atteint généralement qu'avec les amplificateurs stabilisés par relais modulateurs. Cet amplificateur ne comprend donc pas de relais modulateur. Il en résulte moins de bruit et une absence complète de distorsions dues aux intermodulations.

L'amplificateur est également caractérisé par un rejet élevé de mode commun, supérieur à 120 dB, et il n'est pas affecté par une asymétrie de câble d'entrée allant jusqu'à 1 kΩ. Les tensions de mode commun peuvent s'élever à 160 V efficaces. L'un ou l'autre des câbles d'entrée peut être mis à la masse sans affecter le zéro de sortie.

L'atténuateur à 10 plots a un rapport global de 1000 : 1. Une commande de gain intermédiaire supplémentaire de 2,5 : 1 par bouton Vernier est assurée par un potentiomètre à dix tours, muni d'un cadran à verrouillage à divisions de 1 000°.

Le modèle S.E. 4 200 peut être prévu avec un certain nombre d'options:

Sortie ±1,4 V à ±8 mA.
±10 V à 100 mA ou 1 mA.

Le gain de l'amplificateur est de 1 400 ou 10 000 respectivement, c'est à dire une sortie totale pour une entrée de signaux de 1 mV.

La stabilité globale et la précision sont supérieures à ±0,1 % de la déviation sur la totalité de l'échelle pour un réglage d'atténuateur de 10 mV. La réponse de fréquence du système amplificateur est supérieure à ±10 pour cent dans la gamme de 0-10 kHz. Le facteur de bruit est supérieur à 4 µV de crête à crête par rapport à l'entrée.

Des options supplémentaires sont constituées par un système d'étalonnage incorporé pouvant être actionné à distance ou relié à des calculatrices ou à des systèmes de concentration de données.

Chacun des modules amplificateurs mesure 5,08 cm × 16,51 cm de haut sur environ 33,02 cm de profondeur, et on peut loger jusqu'à 8 modules dans un coffret monté sur bâti de 48,26 cm.

EE 93 767 *pour plus amples renseignements*

RÉPARTITEUR DE RÉSISTANCE

Cambion Electronics Products Ltd, Castleton, Nr. Sheffield, Yorkshire

(Illustration à la page 334)

Le "Patch-Ohm" Cambion est un répartiteur de résistance miniature mesurant environ 31 mm de côté et fournissant 56 valeurs de résistance standard de 5 % allant de 51 Ω à 9,5 kΩ.

La résistance "Patch-Ohm" est fabriquée selon des méthodes microélectroniques perfectionnées. Elle se compose d'un réseau de films de résistance d'une grande fiabilité et donnant des résistances standard avec des tolérances de 5 % mises au point par Microtek Electronics (MTE). Le réseau de films épais de résistance est rendu passif par l'emploi du verre suivant un nouveau procédé. Le coefficient de température de la résistance "Patch-Ohm" est de 200 dans 10^6 de −80° C à +150° C. La puissance nominale du matériau de résistance à film épais est de 15 W par pouce carré.

Les différentes valeurs de résistance sont obtenues par une répartition différente entre les huit douilles qui reçoivent les cartes de répartition Cambion "piggy-back" de 0,04 pouce de diamètre.

EE 93 768 *pour plus amples renseignements*

CONTRÔLEUR DE COMPOSANTS

Distributeurs: Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 334)

La Instrument and Digital Measurement Division de la société Dawe Instruments Ltd, en sa qualité d'agent exclusif pour le Royaume-Uni de la société Danbridge du Danemark, est maintenant en mesure de fournir le contrôleur de composants RCL type CPT2 fabriqué par cette compagnie danoise de renommée mondiale.

Cet instrument fournit un pont de comparaison à lecture directe pour comparer les impédances et les angles de phase sur une gamme étendue d'impédances et de fréquences. Il se caractérise par sa grande précision et sa stabilité, ainsi que par sa facilité de fonctionnement.

L'oscillateur RC fournit des fréquences d'alimentation de 100 Hz, 1 kHz, 10 kHz et 100 kHz au circuit de pont et des dispositifs sont prévus pour la lecture directe sur deux instruments de mesure séparés.

Les applications typiques du contrôleur sont les essais de recette, de production et de qualité de composants et de pièces de circuit, la mesure d'inducteurs, de condensateurs ordinaires et de condensateurs électrolytiques, par rapport à des condensateurs superposés de courant continu et à passe-bas. L'instrument est également indiqué pour le contrôle et le réglage de composants groupés.

Les applications de laboratoire comprennent la mesure de la stabilité à long terme, de l'influence de la température et de l'humidité, la comparaison d'impédances de haute précision et la mesure de phases.

La gamme d'impédances va de 3 Ω à 5 MΩ dans la gamme de 0,3 V, de 100 Ω à 5 MΩ dans la gamme de 3 V. Des valeurs de haute impédance atteignant 50 MΩ peuvent être mesurées avec la même précision par un réglage d'étalonnage.

EE 93 769 *pour plus amples renseignements*

RELAIS PHOTOÉLECTRIQUES

O.P. Instruments Ltd, 246 Sutton Common Road, Sutton, Surrey

(Illustration à la page 334)

Dans cette nouvelle gamme de relais et de têtes photoélectriques, tous les éléments sont contenus dans des coffrets en aluminium à l'épreuve des intempéries. Ils sont exclusivement dotés de plaquettes à circuit à fiches remplaçables. La plupart des systèmes peuvent être munis d'un dispositif d'alarme en cas de panne de lampe ou de courant.

En dehors des opérations normales d'arrêt/marche, on peut monter une variété de panneaux de circuit qui assurent un délai avant le fonctionnement du relais ou une fermeture contrôlée et maintenue du relais.

La plupart des relais peuvent être offerts pour le fonctionnement normal d'allumage/extinction ou pour le fonctionnement par impulsion lumineuse momentanée. On peut également prévoir des circuits pour fonctionnement par deux cellules photoélectriques, la lumière sur une des cellules fermant le relais, tandis que la lumière sur l'autre cellule ouvre le relais.

Une variété de têtes de cellule et de lampe est prévue. Les têtes miniatures peuvent, dans des conditions favorables détecter des objets d'un diamètre minimum de 1 mm.

Les éléments sont normalement prévus pour le fonctionnement sur courant secteur mais des systèmes portatifs fonctionnant sur batterie peuvent être fournis.

EE 93 770 *pour plus amples renseignements*

BLOC D'ALIMENTATION STABILISÉE TRÈS HAUTE TENSION

Derritron Instruments Ltd, Parklands, Caincross, Stroud, Gloucestershire

(Illustration à la page 334)

La société Derritron Instruments Ltd présente un bloc d'alimentation e.h.t. au prix le plus bas jamais atteint pour un élément de telle qualité et de telle puissance.

Ce bloc, qui porte le numéro de catalogue E.0411, fournit une tension à variation continue de 500 V à 3 kV jusqu'à 10 mA. Le facteur de stabilisation est de 2000:1 et l'ondulation de crête à crête est limitée à 17 mV à pleine charge. L'impédance de sortie ne dépasse pas 100 Ω.

Le bloc 0411 a été conçu comme élément de laboratoire et il est bâti sur panneau frontal standard de 48,2 cm. Il peut être retiré de son coffret et monté sur bâti, si nécessaire. La sortie est entièrement isolée et elle peut être commutée pour fournir soit une alimentation positive soit une alimentation négative.

Des précautions rigoureuses de sécurité ont été prises pour assurer l'isolement du châssis de toutes les douilles, commutateurs et potentiomètres.

EE 93 771 *pour plus amples renseignements*

AMPLIFICATEUR DIFFÉRENTIEL À GAIN ÉLEVÉ

Tektronix U.K. Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Illustration à la page 335)

Un processus d'équilibrage de tension continue simplifié, une sensibilité étendue et un choix varié de réponses de fréquence font de l'amplificateur type 1A7 un instrument universel à fiches et à gain élevé d'une grande utilité pour l'emploi avec les oscilloscopes de la série Tektronix 530, 540, 550 et 580 (un adaptateur type 81 est nécessaire pour ce dernier).

La gamme totale de sensibilité étalonnée s'étend de 1 μV/cm à 10 V/cm avec largeur de bande de la tension continue à 500 kHz sur toutes les sensibilités.

Le choix de la largeur de bande d'amplification s'effectue à l'aide des deux commutateurs du panneau frontal qui fournissent des points de 3 dB de haute fréquence 8 et de basse fréquence 7.

Le bruit d'entrée est de 3,3 μV efficaces au maximum (dérive non comprise) et la dérive de tension continue ne dépasse pas 200 μV/h de moyenne pendant 10 h avec une température ambiante et une tension de lignes constantes.

Le type 1A7 peut rejeter des signaux de mode communs avec des amplitudes allant jusqu'à 20 V de crête à crête et des fréquences atteignant 100 kHz, avec un rapport de rejet de mode commun de 50 000:1.

L'amplificateur comprend, en outre, un dispositif de compensation différentiel de 300 mV c.c., une lumière de surcharge différentielle et une douille de sortie de signaux.

EE 93 772 *pour plus amples renseignements*

PLAQUETTE À CIRCUIT

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustration à la page 335)

La société A.P.T. Electronic Industries Ltd vient d'ajouter une plaquette à circuit à sa gamme de composants pour le système à châssis modulaire Minar.

La plaquette à circuit Minar, portant la référence C.B.1, est une plaquette isolée au plastic phénolique mesurant 16,3 × 10,8 mm et elle porte un réseau de trous suivant la grille internationale standard de 2,5 cm. N'importe quel circuit électronique peut être fixé sur la plaquette moyennant une perforation réduite ou sans perforation aucune car tous les composants à extrémité de fil métallique peuvent se monter directement sur la plaquette et ils peuvent être reliés entre eux par leur propre fil sur le revers de la plaquette. Ces composants peuvent également être bobinés sur des broches de soudure "Lektrokit" et insérés dans les trous de manière à fournir des bornes à deux extrémités, si nécessaire.

La plaquette de circuit a été conçue de façon à pouvoir être montée sur des supports derrière les panneaux frontaux à module individuel, chaque module pouvant porter une ou deux plaquettes.

Le système de construction modulaire Minar, dérivant du système à châssis MINiature, a été conçu par le service de recherches de la British Broadcasting Corporation pour l'emploi dans la construction d'appareils de télévision en couleurs et stéréo expérimentaux et il est construit et distribué par A.P.T. sous forme de système à châssis durable et d'une grande souplesse d'emploi pour la construction de circuits expérimentaux.

EE 93 773 *pour plus amples renseignements*

LIGNE À RETARD DE TÉLÉVISION

Lexor Electronics Ltd, 25-31 Allesley Road, Coventry

(Illustration à la page 335)

Prévoyant la demande future de composants spécialisés pour la télévision en couleurs, la société Lexor a mis au point une ligne à retard à bas prix et ayant la fonction d'un filtre passe-bas.

Cette ligne sert de filtre pour la correction de phase du canal de luminance par rapport aux canaux de chrominance. Les paramètres précis dépendent du type de système de couleurs et de circuit utilisés.

Un élément-type, tel qu'on le voit dans notre gravure, comporte un retard de 800 nsec à une impédance de 1,8 kΩ et une fréquence de coupure de 2,5 MHz avec réponse de phase linéaire. Il mesure 15,24 cm de longueur × 1,58 cm de diamètre et il est sous encapsulage de matière plastique.

Si ces éléments sont commandés en quantitées suffisantes, leur prix peut être extrêmement bas.

EE 93 774 *pour plus amples renseignements*

INSTRUMENT DE MESURE DE MODULATION D'AMPLITUDE/ MODULATION DE FRÉQUENCE

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustration à la page 335)

La société Marconi Instruments Ltd, vient d'annoncer la sortie d'un instrument universel "tout transistors" pour la mesure de la modulation. Cet instrument, type TF 2300, a été conçu pour répondre à toutes les utilisations dans le domaine des communications fixes et mobiles de "point à point," ainsi que pour la plupart des applications de radiodiffusion et de télémétrie.

La gamme de fréquence porteuse s'étend de 4 MHz à 1 000 MHz. Cinq gammes de déviation, de 5 kHz à 500 kHz, sont prévues avec une gamme de modulation de fréquence de 30 Hz à 150 kHz. Pour les mesures dans la bande acoustique, la gamme de fréquence de modulation peut être réduite à 15 kHz au moyen d'une commande de panneau.

Deux gammes de mesure en profondeur de la modulation d'amplitude sont prévues, soit 0 à 30 % et 0 à 100 % aux fréquences de modulation fondamentale dans la gamme de 30 Hz à 15 kHz. La gamme de modulation de fréquence s'étend à 50 kHz pour la mesure de distorsions.

Il existe une sortie de signaux démodulés à un niveau de 1 mW dans 600 Ω pour une déviation totale de l'échelle d'un instrument de mesure, tant pour les signaux de modulation de fréquence que pour les signaux de modulation d'amplitude. Un filtre standard d'atténuation de 50 µsec ou 75 µsec peut être branché au circuit de sortie de signaux démodulés pour la mesure sur les émetteurs à "préemphase."

L'oscillateur local permet le contrôle piloté par quartz de la fréquence pour éliminer les bruits de l'oscillateur local et pour permettre le contrôle de production rapide jusqu'à trois fréquences porteuses prédéterminées. Il existe également un dispositif d'entrée externe de l'oscillateur local. Une source d'étalonnage à cristal incorporée permet de normaliser l'indication de déviations.

L'instrument fonctionne sur tension secteur normale ou sur alimentation à la terre positive en tension continue nominale de 24 volts.

EE 93 775 *pour plus amples renseignements*

INSTRUMENT DE MESURE DES INTERVALLES DE TEMPS

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustration à la page 335)

L'instrument de mesure des intervalles de temps, type 369, est un appareil peu coûteux et entièrement transistorisé pour la mesure d'augmentation de temps très réduite dans la gamme de 100 nsec à 30 msec.

Il s'agit essentiellement d'un chronomètre électronique de grande précision utilisant des méthodes de comptage en liaison avec des oscillateurs intérieurs stables. L'instrument fonctionne à partir de plots ou impulsions positives ou négatives et il peut être réglé pour l'utilisation à n'importe quel niveau de 1 V à 100 V; on peut utiliser soit le fonctionnement sur une ligne soit sur deux lignes.

Cette combinaison de possibilités permet d'effectuer la mesure de la largeur d'impulsions, de l'intervalle d'impulsions, des temps de montée ou de chute, ainsi que la mesure de l'intervalle de temps entre des impulsions sur deux lignes différentes.

L'indication est donnée sur un instrument de mesure à échelle ouverte de 12,7 cm, la lecture étant maintenue indéfiniment jusqu'au réenclenchement.

EE 93 776 *pour plus amples renseignements*

Résumés des Principaux Articles

Les principes électroniques du Mekometer II par R. H. Bradsell

Résumé de l'article aux pages 282 à 287

Le Mekometer II est un instrument électronique de mesure de la distance prévu pour mesurer des distances allant jusqu'à 1,5km de la plus simple manière possible.

Il utilise un rayon lumineux modulé à près de 500MHz. Des résonateurs à cavité spécialement conçus déterminent la longueur d'onde de modulation, indépendamment de la connaissance de la vitesse de la lumière et, par conséquent, de l'indice réfractif de l'air le long du parcours du rayon lumineux.

Un circuit de pont R-C perfectionné à haute sélectivité par S. Bernstein

Résumé de l'article aux pages 288 à 293

L'auteur présente un circuit de pont R-C dont le fait de lui avoir ajouté un étage unité-gain actif a eu pour résultat d'augmenter la valeur Q de circuit effectif à une valeur représentant 312 de la valeur obtenue avec un pont de Wien classique. Cette amélioration est particulièrement utile dans les circuits d'amplificateurs sélectifs et d'oscillateurs fonctionnant aux fréquences acoustiques ou subacoustiques.

Circuits de modulation à impulsions à pente variable et à transition arrière par O. E. Kruse et J. K. Kincaid

Résumé de l'article aux pages 294 à 295

Les circuits électroniques pour la production et la détection d'un train d'impulsions modulées triangulaires sont décrits dans cet article. L'information devant être transmise est portée sur les pentes des flancs arrière du train d'impulsions. Dans le processus de démodulation, les impulsions de pente variable sont convertis en impulsions à amplitude variable, ces dernières étant ensuite amplifiées, étendues et filtrées. La sortie finale constitue une bonne reproduction de l'information originale.

Un synthétiseur de fréquence d'une grande souplesse d'utilisation prévu pour les radiocommunications mobiles
par A. F. Evers

Résumé de l'article aux pages 296 à 303

Cet article décrit une source de fréquence accordable de precision utilisant un oscillateur à fréquence variable et à asservissement de phase. Le système de contrôle de fréquence à circuit fermé, qui comprend des diviseurs de fréquence à rapport variable et à commande numérique, compare un sous-multiple variable de la fréquence de sortie de l'oscillateur à fréquence variable avec un sous-multiple fixe de la sortie d'un étalon de fréquence à stabilité élevée. Ce synthétiseur à fréquence numérique conviendrait à une gamme étendue d'applications dans les appareils mobiles de radiocommunications où un encombrement réduit, une faible consommation de courant et une haute fiabilité sont des considérations d'importance. On peut obtenir une sortie extrêmement pure moyennant une consommation de courant légèrement supérieure et un appareil un peu plus complexe.

L'auteur analyse certains des problèmes de réalisation les plus importants et décrit les caractéristiques de plusieurs prototypes de laboratoire et de production du synthétiseur numérique.

Un point sur un commutateur de cycles par R. A. Stevens

Résumé de l'article aux pages 304 à 306

Dans de nombreux cas, lorsqu'on étudie les conditions transitoires dans les courants à tension alternative, il est nécessaire de pouvoir brancher à un point prédéterminé de l'onde d'alimentation et de répéter le point de commutation avec précision. On a déjà construit des appareils pour effectuer cette opération mais on a décidé cette fois de réaliser un système entièrement constitué de corps solides. Ce système utilise une méthode de déphasage assez bien établie qui comporte le fonctionnement d'une porte à un niveau choisi sur une rampe de tension. Le dispositif est compact et portatif et il est très utile pour les travaux d'études universitaires en laboratoire, en raison surtout de son mode de fonctionnement à répétition.

Un système modulateur/démodulateur de périodes par L. Molyneux

Résumé de l'article aux pages 307 à 309

L'auteur décrit le circuit ainsi que le comportement d'un système modulateur/démodulateur de périodes. Le principal avantage du système par rapport à la modulation de fréquence, à laquelle il ressemble étroitement, réside dans la simplicité du filtre passe-bas qui suit le démodulateur; dans certains cas, le filtre peut être entièrement supprimé.

Un circuit transistorisé d'entraînement par impulsions pour un moteur à échelons par W. J. Jirafe

Résumé de l'article aux pages 316 à 321

Cet article décrit la mise au point d'un circuit transistorisé servant à entraîner un moteur à échelons standard à quatre conducteurs soit dans le sens des aiguilles d'une montre soit dans le sens inverse par des impulsions à son entrée. Le moteur avance d'un échelon par impulsion et la séquence peut être contrôlée au moyen d'un commutateur. La séquence de commutation est produite par deux multivibrateurs dont les sorties amplifiées entraînent les deux phases du moteur. Elles sont déclenchées par un multivibrateur commun actionné par des impulsions d'entrée à travers une logique de séquence. Le circuit est à protection automatique contre les dérangements de secteur et autres. Un dispositif différent, utilisant un compteur binaire réversible, est proposé; les caractéristiques favorables du dispositif précédent par rapport à ce dernier sont mises en évidence. L'article indique également le montage prévu pour un moteur avec enroulements bifilaires. Une brève étude du mode de fonctionnement d'un moteur à échelons et son utilisation comme servomécanisme de position est également effectuée.

L'étude analytique d'un réseau à termistance pour la stabilisation en continu de transistors par H. Singh

Résumé de l'article aux pages 322 à 324

Une méthode analytique a été développée pour l'étude d'un réseau à termistance pour la stabilisation du point de fonctionnement en continu de transistors. Cette méthode est basée sur le principe de la technique de compensation de thermistance qui permet de stabiliser le courant au collecteur en réglant la tension base/émetteur du transistor à trois valeurs de température presélectionnées. La méthode dont il est question ici améliore considérablement la stabilisation du courant continu au collecteur dans la gamme de température de 20–50°C.

Cette méthode sert de variante pour former les caractéristiques de thermistance suivant les méthodes graphiques.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 330 bis 335

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Phasenempfindlicher Detektor

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(*Abbildung Seite 330*)

Fertigung eines neuen transistorisierten, phasenempfindlichen Detektors für den Frequenzbereich 10 Hz ... 300 kHz mit eingebautem Anzeigegerät und Ausgangssignalen, die eine breite Auswahl von Schreibern, Digitalvoltmetern usw. treiben können, wird von Brookdeal Electronics Ltd angekündigt.

Hochgradige Linearität und niedrige Nullpunktdrift ermöglichen Verwendung des Instruments bei sehr kleinen Signalpegeln in Gegenwart viel grösserer Rauschspannungen; es ist als Wechselstromdetektor in Präzisionssystemen besonders nützlich. Der Verringerung der mit der Einstellung verbundenen Vorgänge wurde besondere Aufmerksamkeit gewidmet und als Resultat der Frequenzbereichschalter und die Bezugsspannung eliminiert.

Zur Überwachungsschaltung gehören ein Grossskaleninstrument mit Nullpunkt in der Mitte oder links sowie umfassende Einrichtungen zum Einschalten von Zeitkonstanten zwischen 0,03 und 3 Sekunden, Drucktastenabschwächer und auf Wunsch ein Filter für niederfrequenten Betrieb sowie Elemente zur Nullpunktunterdrückung.

Das Gerät ist für Einbau in 19"-Gestelle ausgelegt, kann aber auch in einem Instrumentgehäuse oder mit einem geeigneten Vorverstärker in einem Doppelgehäuse geliefert werden.

EE 93 751 *für weitere Einzelheiten*

Relais

Painton & Co. Ltd, Kingsthorpe, Northampton

(*Abbildung Seite 330*)

Ein Lizenzvertrag mit Chauvin Arnoux in Paris gibt Painton & Co Ltd die ausschliessliche Herstellung der Relais-Baureihe OK für Grossbritannien. Painton hat ausschliessliche Vertriebsrechte für diese Relais nicht nur in Grossbritannien, sondern auch in Irland, Australien, Neuseeland, Kanada und Südafrika.

Die Relais haben einen Höchstnennstrom von 10 A pro Kontakt, sind mit vier Umschaltkontakten bestückt und haben eine lange Lebensdauer von mehreren Millionen Schaltspielen. Sie haben breite Anwendungsmöglichkeiten, und verschiedene Modelle sind für Schütze, Zeitverzögerung und Logik sowie Sonderzwecke lieferbar.

Dank neuer Konstruktionsmethoden ist die Justierung des Relais permanent und sowohl in magnetischer wie auch mechanischer Beziehung äusserst zuverlässig. Alle Abwandlungen dieser Relais werden in ein staubdichtes, transparentes Gehäuse mit kompakten Abmessungen eingebaut. Sowohl Gleichstrom- wie Wechselstromausführungen sind lieferbar. Ein breites Sortiment von Montagezubehör ist vorhanden; die Sockelstifte sind so ausgeführt, dass sie für Montage in Fassungen oder mit Ansteckverbindungen geeignet sind.

EE 93 752 *für weitere Einzelheiten*

Digitalvoltmeter

Vertrieb: Kynmore Engineering Co. Ltd, Buckingham Street, London, W.C.2

(*Abbildung Seite 330*)

Ein neues Digitalvoltmeter mit 0,1 Prozent Messgenauigkeit, in modularer Bauart und ergonomisch gestaltet, ist nunmehr von Kynmore Engineering Co Ltd erhältlich. Es ist das von der französischen Firma Schneider hergestellte Digitalvoltmeter Venus und kommt in zwei Standardausführungen, jedes mit einer Auswahl von Verfeinerungen, durch die es zusammen mit einem Zubehör-Sortiment für hohe Ansprüche stellende Messzwecke geeignet ist.

Das eine Standardmodell—Venus 123—hat drei Messbereiche: $-9{,}99 \ldots +9{,}99$ V; $-99{,}9 \ldots +99{,}9$ V; $-999 \ldots +999$ V. Kommastellung und Vorzeichenanzeige sind automatisch, und die Messunsicherheit ist 5×10^{-4} des Bereichendwertes ± 1 Ziffer.

Das andere Modell ist Venus 124 mit einer zusätzlichen Einzelziffer für Überlastungsanzeige. Die jeweiligen Höchstwerte für die drei Messbereiche sind daher 10,00, 100,0 und 1 000 V; es werden aber höhere Werte angedeutet.

Mit Ausnahme des 1 000-V-Bereiches sind Überlastungen von etwa 25 Prozent für beide Modelle zulässig. Die Mindesteingangsimpedanz ist 20 MΩ, mit 1 GΩ für den 10-V-Bereich gegen Aufschlag. Der Betriebstemperaturbereich ist $+5 \ldots +45°$ C. Die Messungen werden automatisch jede Sekunde wiederholt.

Wahlweise Sondereinrichtungen sind unter anderem ein Druckwerk—beide Modelle können mit dem ebenfalls von Kynmore lieferbaren Schneider-Drucker arbeiten—oder der Schneider-Zubehör Mars für Umwandlung in ein Digital-Mehrfachmessgerät. Der Mars-Zubehör gibt zwei niedrige Gleichspannungsbereiche bis zu 100 mV Vollausschlag herunter, vier Wechselspannungsbereiche von 300 mV_{eff} bis zu 300 V_{eff} Bereichendwert, sowie Widerstandsbereiche. Messunsicherheit, Empfindlichkeit und Eingangsimpedanz hängen vom benutzten Bereich ab, sind aber mit den mit dem Venus allein erzielbaren vergleichbar.

Venus arbeitet mit Pulslängenmodulation und ist mit einem quarzoszillatorgesteuerten PDM-Differentialdemodulator ausgerüstet. Automatische Polaritäts- und Spannungsanzeige werden durch interne Logik erreicht, und das Resultat wird mit Hilfe von Kaltkathodenanzeigeröhren dargestellt. Bereich oder Eichung werden mittels Drucktasten gewählt, die Bezugsspannung aber intern eingeregelt. Nach der ersten Einstellung bleibt sie nach der Anheizzeit von 10 Minuten für lange Zeit konstant.

EE 93 753 *für weitere Einzelheiten*

Keramische Scheiben-Trimmerkondensatoren

Steatite Insulations Ltd, Hagley House, Edgbaston, Birmingham, 16

(*Abbildung Seite 331*)

Steatite Insulations Ltd hat ihr Programm für keramische Subminiatur-Scheibentrimmer weiter ergänzt.

Links in der Abbildung wird Typ 7 S-Triko 03 gezeigt, der den anspruchsvollen Anforderungen der Anwender für militärische und professionelle Zwecke genügt. Der Trimmer entspricht den Vorschriften MIL-C 81A und ist mit folgenden Kapazitätsbereichen lieferbar:

P 100	2 ... 3,5	pF
N 033	2,5 ... 6	pF
N 075	3 ... 10	pF
N 470	3,5 ... 13	pF
N 750	4,5 ... 20	pF
N1500	7 ... 35	pF

Rechts ist Typ 5 S-Triko 03 abgebildet, eine Reihe sehr zuverlässiger keramischer Scheibentrimmer, die mit folgenden Kapazitätsbereichen lieferbar sind:

P 100	1,5 ... 2,5	pF
N 033	2,5 ... 6	pF
N 075	3 ... 8	pF
N 470	3,5 ... 11	pF
N 750	4,5 ... 15	pF
N1500	6 ... 22	pF

EE 93 754 *für weitere Einzelheiten*

Digitalstoppuhr

Darang Electronics Ltd, Restmor Way, Hackbridge Road, Hackbridge, Surrey

(Abbildung Seite 331)

Modell 661D ist eins der neuen “Digicron”-Modelle elektronischer Digitalstoppuhren von Darang Electronics Ltd. Die Konstruktion beruht auf einem System preisgünstiger Einsteckplatten und der äusserst zuverlässigen Kaltkathodentechnik. Zur Befriedigung der Sonderwünsche von Kunden sind viele Ausführungen möglich.

Gasgefüllte Ziffernröhren geben eine einzeilige, fünfstellige Digitalanzeige. Die überstrichene Zeitspanne ist 59 Minuten 59,9 Sekunden in Schritten von 0,1 Sekunden. Im Standardmodell dient die Netzfrequenz als Zeitquelle, man kann aber auch einen externen Zeitgeber benutzen. Gegen Zuschlag kann ein Zeitnormal mit Kristall in einem Thermostaten eingebaut werden.

Stopp und Start werden durch Drucktasten auf der Frontplatte oder durch externe Fernkontakte gesteuert. Die Kabellänge zu den externen Kontakten kann über 1,6 km sein.

Rückstellung ist bei Betätigung der Start-Taste automatisch; es ist jedoch auch möglich, mehrere Zeitperioden zu addieren und die Gesamtzeit anzuzeigen.

Das Gerät ist für Anschluss an 95... 130 V oder 190...260 V, 50 Hz ausgelegt.

EE 93 755 *für weitere Einzelheiten*

Galvanische Hautwiderstandsmessungen

Short Brothers & Harland Ltd., Precision Engineering Division, Castlereagh, Belfast, N. Ireland

(Abbildung Seite 331)

Das Short-Messgerät G.S.R. ist ein in sich geschlossenes Instrument mit Netzanschluss für 230 V~, das Durchführung folgender Messungen gestattet:

(a) Hautwiderstand eines Patienten

(b) Widerstandsänderungen als Reaktion auf verschiedene Reize, die entweder Angstzustände hervorrufen oder neutral wirken.

Die Zuleitungen der Hautelektroden werden in Buchsen auf der Frontplatte gestöpselt, und durch den zu testenden Hautweg wird ein konstanter Strom von 10 μA gespeist. Ein Frontplatteninstrument zeigt den Basiswiderstand an, wird aber durch Widerstandsänderungen nicht beeinflusst. Das Gerät hat drei Messbereiche mit Endwerten von 100 kΩ, 200 kΩ und 500 kΩ. Ein über eine Frontplattenklinke angeschlossener Schreiber registriert nur die Widerstandsänderungen und schreibt null, wenn keine Änderungen auftreten.

Die Trennung des Basiswiderstandes und der Widerstandsänderungen gibt dem Gerät einen grossen Vorteil über viele andere Geräte dieser Art, bei denen der Bedienende durch Regelung eines Potentiometers für den Basiswiderstand kompensieren muss, um die Änderung zu beobachten. Die einzige Einregelung, die das Short-Gerät erfordert, ist tägliche Nullung des Verstärkers von der Frontplatte aus.

EE 93 756 *für weitere Einzelheiten*

Impulsgeber

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Abbildung Seite 331)

Dynamco Instruments Ltd hat nunmehr die Fertigung von zwei neuen Ausführungen des Festkörper-Impulsgebers Datapulse 110 in England aufgenommen. Es handelt sich um Modell 110A, das ohne Preiserhöhung eine verbesserte Anstiegszeit und externe Triggerung gibt, sowie Modell 110P, eine programmierte Ausführung des 110A.

Modell 110A bietet eine Minimumanstiegszeit von 4,5 ns (typisch 4 ns) und externe Triggerung von 1 V ab. Die Wiederholungsfrequenz ist zwischen 4 Hz und 40 MHz regelbar. Die linearen Sprungzeiten sind getrennt regelbar. Gleichzeitige positive oder negative Ausgangsimpulse werden bis zu 10 V an 50 Ω mit bis zu 70 dB Abschwächung entweder in Einzel- oder Doppelimpulsbetrieb abgegeben. Die Impulsdauer ist zwischen 10 ns und 5 ms, die Laufzeitkette zwischen 10 ns voreilend und 50 ms Verzögerung regelbar. Ein typisches Tastverhältnis ist 85 Prozent. Eine veränderliche Gleichstrom-Grundlinie erlaubt die Wahl von Komplementärausgangsimpulsen bei jedem Pegel zwischen ± 10 V. Grossbereichregler gestatten Prüfung auf Toleranz, erleichtern Wiederholbarkeit der Testergebnisse und ermöglichen genaue Nachbildung von Sägezahn- und Dreieckwellenformen für Impulsmessungen der Induktivität und Verstärkertests. Einzelausgangsimpulse gestatten Fehlersuchen an Logikschaltungen. Keine Kombination der Reglereinstellungen auf der Frontplatte, noch Kurzschlüsse irgendwelcher Dauer oder Sperrspannungen bis zu ± 10 V können die Ausgangszustände schädigen. Das Gerät ist für Betrieb bei Umgebungstemperaturen von 0...50° C bemessen.

Der programmierte Impulsgeber 110P ermöglicht Fernsteuerung und/oder automatisches Programmieren aller Impulsparameter einschliesslich Anstiegszeit und Verschiebung der Gleichstrom-Basislinie. Einsatz mit Zubehör und Lochkarten- oder Lochstreifenlesern oder Magnetscheibenprogrammieren ermöglicht schnelles Testen von Bauelementen wie z.B. integrierte Schaltungen.

EE 93 757 *für weitere Einzelheiten*

Phasenwinkel-Voltmeter

Smiths Industries Ltd. Aviation Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Abbildung Seite 331)

Ein neues kompaktes 400-Hz-Festkörper-Phasenwinkel-Voltmeter mit vielen nützlichen Eigenschaften wird vom Geschäftsbereich Luftfahrt der Smiths Industries Ltd angeboten. Bei 216 mm Länge, 152 mm Höhe und 203 mm Tiefe hat das Gerät eine Messunsicherheit von ± 3 Prozent für Amplituden und ± 1 Grad für Phasen.

Das Gerät kann entweder die aufgelösten Komponenten oder Phase und Eingangsmodulus darstellen und hat 12 Empfindlichkeitsbereiche mit Endwerten zwischen 300 V und 1 mV. Das Bezugssignal kann man mit beliebigem Wert zwischen 300 V und 100 mV anlegen; eine selbsttätige Verstärkungsregelung macht Pegeleinregelung von Hand überflüssig. Das einzige Einstellorgan braucht nur bei Änderung der Signalfrequenz nachgestellt zu werden. Sowohl Signal- wie Bezugseingang sind transformatorgekoppelt und geben die für gültige Brücken- und Differentialmessungen erforderliche Gleichtaktunterdrückung von 10^5:1. Sowohl im Signal- wie im Bezugskanal eingebaute Filter sorgen für die notwendige Unterdrückung von Harmonischen.

Ein Überlastungsfaktor 10:1 wurde in Rechnung gestellt, so dass kleinere Komponente in Gegenwart eines grossen Gesamtsignals gemessen werden können. Eine Signallampe leuchtet auf, wenn der Faktor überschritten wird.

Man kann das Gerät an jede Stromversorgung zwischen 100 V und 250 V, 50...400 Hz anschliessen. Bei Änderung der Spannung oder Frequenz ist keine Umstellung erforderlich, so dass beim Anschliessen kein Fehler möglich ist, wenn mehr als eine Stromversorgung vorhanden ist.

EE 93 758 *für weitere Einzelheiten*

Transistorisiertes Sende-Empfangsgerät

G.E.C. (Electronics) Ltd, Communications Group, Spon Street, Coventry

(Abbildung Seite 332)

Das Modell RC/420/TR wiegt nur 6,8 kg und ist für Sprech- oder CW-Weitverkehr im HF-Band zwischen 2 und 18 MHz ausgelegt. Es kann als feste oder bewegliche Station eingesetzt werden und passt ohne Schwierigkeiten unter das Armaturenbrett eines Kraftfahrzeugs.

Das Sende-Empfangsgerät verkörpert einige interessante Konstruktionsmerkmale. Die meisten Schaltungen sind mit einem einzigen Silizium-Mesatransistortyp aufgebaut. Germaniumtransistoren gibt es nur in der Stromversorgung und den Niederfrequenzstufen; das verbessert Zuverlässigkeit und Wartung. Vier quarzgesteuerte Kanäle im 2 ... 18-MHz-Band können im ESB- oder 2SB-Betrieb arbeiten. Bei Einbau eines wahlweisen Filters für das untere Seitenband kann das Gerät mit acht Frequenzkanälen betrieben werden. Eine vom Personal bediente "Klär"-Schaltung verringert den Bedarf an Thermostaten mit hoher Leistungsaufnahme und damit die Stromentnahme von der Batterie. Alle Schaltungskarten und Hauptteile sind gegen Pilzwuchs behandelt.

Das Gerät wird komplett mit elektrodynamischem Mikrofon geliefert. Ein breites Zubehör-Sortiment ist lieferbar, einschliesslich einer Einrichtung für direkten Verkehr mit Fernsprechleitungen, einer Stromversorgung für 110/220 V~, einer Antennenabstimmung für Mehrfrequenzeinsatz und einem Gestell sowie Antenneninstallation für bewegliche Verwendung. Ein herkömmlicher Handsprechhörer sowie ein Fernschreiberendapparat und ein Druckwerk sind ebenfalls lieferbar.

EE 93 759 *für weitere Einzelheiten*

Messgerät für Klirrfaktor, Dezibel und Millivolts

Laboratoire Electro-Acoustique, 5 Rue Jules Parent, Rueil (S & O), Frankreich

(Abbildung Seite 332)

"Qualiscope" ist ein Mehrzweck-Präzisionsinstrument, das in einem Gerät die Mittel zum Messen der Haupteigenschaften einer Schaltung enthält. Die gemessenen Parameter sind: Klirrfaktor (in Prozent), Spannung (mV), Spannungspegel (dB) und Rauschpegel.

Der Klirrfaktormesser überstreicht 10 Hz ... 100 kHz mit Grundfrequenzen und 10 Hz ... 250 kHz mit Harmonischen; die Auflösung ist 0,1 Prozent des Vollausschlags.

Der Spannungs- und Dezibelmesser hat einen Frequenzbereich von 2 Hz ... 400 kHz mit einem Frequenzgang innerhalb ±0,2 dB von 20 Hz ... 100 kHz und ±2 dB von 2 Hz ... 400 kHz. Der Spannungsmessumfang ist 1 mV ... 300 V Endwert. Ein zusätzlicher Vorverstärker (Verstärkung +20 dB) erhöht die Empfindlichkeit auf 100 µV Vollausschlag.

Das Gerät hat eine eingebaute Oszillografenröhre zur Darstellung Lissajousscher Figuren.

EE 93 760 *für weitere Einzelheiten*

Automatischer Untersetzer

L'Electronique Appliquée, 96-100 Rue Maurice Arnoux, Montrouge (Seine), Frankreich

(Abbildung Seite 332)

Der transistorisierte Untersetzer hat für die Anzeige sechs Nixie-Röhren und einen sechsstelligen elektromechanischen Zähler mit Drucktastenrückstellung. Die Betriebsart wird mittels Drucktasten auf der Frontplatte eingestellt.

Der Untersetzer zählt Impulse mit einer zwischen 0,1 V und 10 V einstellbaren Empfindlichkeit; die Eingangsauflösung ist 1 µs. Die Eingangsimpedanz ist 10 kΩ für positive Impulse und 2,5 kΩ für negative.

Bei Frequenzmessungen wird während einer Sekunde gezählt und für die nachfolgenden zwei Sekunden angezeigt.

Der Untersetzer ist automatisch nach Zählen bis zu einer beliebigen vorgewählten Zahl bis zu $n10^p$ abschaltbar, wobei n zwischen 1 und 9 und p zwischen 0 und 6 einstellbar ist. Andererseits kann der Untersetzer nach Erreichung einer vorgewählten Zeit zwischen 1 und 10^6 Sekunden automatisch aufhören zu zählen. Wenn sowohl Zeit wie auch Zählung vorgewählt sind, stoppt das Gerät, nachdem der erste Wert erreicht ist.

Ein 100-kHz-Quarzkristall dient als Frequenznormal.

Der Ausgang kann ein Druckwerk treiben.

EE 93 761 *für weitere Einzelheiten*

Stellmotor

Servomex Controls Ltd, Crowborough, Sussex

(Abbildung Seite 333)

Dieser neue kleine zweiphasige 50-Hz-Stellmotor mit Tachogenerator und Untersetzungsgetriebe ist einfach konstruiert und äusserst preisgünstig, da er nicht wie andere Stellmotoren vorwiegend für militärische Anwendungszwecke mit der geforderten hervorragenden Leistung bei entsprechend hohen Kosten entwickelt wurde.

Trotzdem entsprechen Werkstoffe und Verarbeitung des neuen Servomex-Motors den höchsten Ansprüchen, so dass er für Einbau in Instrument-Regeleinrichtungen bester Qualität geeignet ist.

Der Motor ist vierpolig mit abgeschrägten Rotorstäben und einem Schleppschalen-Tachogenerator. Lieferbar sind zwei Modelle—SM95 und SM135—die zwar verschiedene Wicklungen und Lager, im Grunde aber dieselben Eigenschaften haben. Die Ausgangswelle des SM95 läuft in porösen Bronzelagern, die mit Silikonschmiermittel imprägniert sind. Die Ausgangswelle des SM135 läuft in einem Präzisionskugellager und hat einen genau bearbeiteten Zentrieransatz.

Die Spulen sind nach dem Pflichtenblatt RCS 214 vakuumimprägniert. Der Getriebekasten hat eine Untersetzung von 40 : 1 über Nylon- und Stahlzahnräder mit Rutschkupplung, die bei etwa 35...40 mg rutscht, damit keine Beschädigung auftritt.

Die Ausgangswelle hat im Leerlauf eine Höchstgeschwindigkeit von 60 UPM und ein Kurzschlussdrehmoment von 21 mg. Beide Modelle werden in ein Gussteil aus Aluminiumlegierung eingebaut und sind für Betriebstemperaturen von −10° C...+70° C geeignet.

Modell SM95 ist für Treiben durch einen Verstärker mit eingebauter Phasenänderung von 90° ausgelegt; sowohl die Motorbezugswicklungen wie die Generatorerregerwicklungen werden mit 6,3 V~ bei Nullphase gespeist. Die Motorbezugswicklung des SM135 ist für Speisung mit 230 V_{eff} durch einen Kondensator von 1 µF...1,4 µF ausgelegt. Dadurch eilt die Wicklung der Speisespannung um 90° voraus und gestattet Erregung der Steuerwicklungen durch einen Verstärker mit Phasenverschiebung von null.

Die Abmessungen sind 76 × 114 × 60 mm, das Gewicht 985 g.

EE 93 762 *für weitere Einzelheiten*

Keramik-Scheibenkondensatoren

Centralab Ltd, Northway House, Great North Road, London, N.20

Es handelt sich um ein preisgünstiges Angebot von Keramik-Scheibenkondensatoren guter Qualität und mit vorteilhaften Abmessungen. Die drei Kapazitätswerte sind 0,01 µF, 0,02 µF und 0,05 µF mit Scheibendurchmessern von 2,5 mm, 9,1 mm bzw. 13,1 mm und nicht dicker als 3,2 mm.

Die Betriebsspannung der Serie ist 50 V−. Mit Ausnahme des grössten Kondensators sind die Anschlussabstände 6,35 mm, für den grössten 9,5 mm. Anschlussdrähte werden entweder gerade oder mit Kink und für Bestückung gedruckter Schaltungen geliefert. Alle Anschlussdrähte sind zwecks leichterer Bestückung verzinnt.

Jedes Bauelement dieses Sortiments wird mit Toleranzen von +80 Prozent bis −20 Prozent oder ±20 Prozent geliefert.

EE 93 763 *für weitere Einzelheiten*

Temperaturregler

Société D'Etudes et Recherches Electroniques, 15 Rue Villebois-Mareuil, 92-Colombes (Seine), France

(Abbildung Seite 333)

Der neue SERELEC-Temperaturregler kombiniert die proportional-, differential-und integralwirkende Regelung und hält sie innerhalb enger Grenzen (typisch 0,2°C/500°C), wobei Schwankungen der Netzspannung, der Umgebungstemperatur oder externe Störungen keinen Einfluss haben.

Er besteht aus drei magnetischen Verstärkerstufen und zwei Thyristoren für die Regelung des Ausgangs.

EE 93 764 *für weitere Einzelheiten*

Elektronisch abstimmbare Filter

Microwave Associates International Inc.: Europäisches Büro: Cradock Road, Luton, Bedfordshire

(Abbildung Seite 333)

Microwave Associates haben für den Frequenzbereich 125 ... 12 000 MHz eine neue Serie elektronisch abstimmbarer YEG-Filter eingeführt. Diese Filter haben einen sehr niedrigen Einfügungsverlust, gute Unterdrückung von Nachbarbändern und sind frei von Unselektivität.

Sie wiegen nur etwa 770 g, haben ein Höchststehwellenverhältnis von 1,5 und entsprechen allen Anforderungen von MIL-E-5400 Klasse II.

Die Filter sind hochpolierte, einkristalline Kugeln aus Yttrium-Eisen-Granat für Anwendung als gyromagnetische Abstimmelemente.

Treiberverstärker sind lieferbar, die mittels eines 0 ... 10-V-Signals von einer hochohmigen Quelle in ihren entsprechenden Bereichen vom unteren bis zum oberen Ende treiben. Wo erforderlich, können auch Stromversorgungen und Temperaturregler geliefert werden.

Diese Filter wurden für Vor- oder Nachselektion mit jedem Empfängertyp entwickelt, der sehr schnelles Abtasten und/oder automatische Signalselektion oder Signalverarbeitung erfordert.

EE 93 765 *für weitere Einzelheiten*

Trimmerkondensatoren

Steatite Insulations Ltd, Hagley House, Hagley Road, Edgbaston, Birmingham, 16

(Abbildung Seite 333)

Das Programm der Steatite Insulations Ltd für Trimmerkondensatoren wurde vor kurzem durch den Schraubentrimmer 1250b erweitert, der für Montage auf gedruckte Schaltungen ausgelegt und mit Höchstkapazität von 10 pF oder 20 pF lieferbar ist; die Mindestkapazität ist in beiden Fällen 2 pF. Der Kapazitätsverlauf ist über einen Trimmbereich von 6 × 360° linear; der Verlustfaktor ist bei 1 MHz besser als 0,002 und die Betriebsspannung 250 V—. Die Aussenabmessungen sind 15,5 mm × 7,5 mm.

EE 93 766 *für weitere Einzelheiten*

Gleichstromverstärker

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Abbildung Seite 333)

S.E. Laboratories (Engineering) Ltd hat einen geräuscharmen Breitband-Gleichstromverstärker S.E. 4200 mit hoher Verstärkung und völlig erdfreiem Differentialeingang für Oszillografen und magnetische Datenspeicherung, moderne Datenerfassungssysteme usw. herausgebracht.

Der S.E. 4200 ist für Verstärkung von Signalen mit sehr niedrigem Pegel geeignet, wie sie von Thermoelementen, Dehnmessstreifen und ähnlichen Elementen abgegeben werden.

Eine seiner Eigenschaften ist die üblicherweise nur mit chopperstabilisierten Verstärkern erreichbare hohe Konstanz, trotzdem es in diesem Verstärker keinen Chopper gibt. Andere Vorteile sind Rauschreduktion und völlige Abwesenheit von Verzerrungen durch Zwischenmodulation.

Die hohe Gegentaktunterdrückung von über 120 dB ist eine weitere Eigenschaft; Unsymmetrie der Eingangszuleitungen bis zu 1 kΩ haben keinen Einfluss. Gegentaktspannungen können bis zu 160 V_{eff} sein. Jede der Eingangszuleitungen kann ohne Einwirkung auf den Ausgangsnullpunkt geerdet werden.

Ein 10stufiger Abschwächer hat ein Gesamtverhältnis von 1 000 : 1. Ein 10-gängiges Potentiometer mit feststellbarer, in 1 000° unterteilter Skala gestattet Zwischenfeinregelung der Verstärkung 2,5 : 1.

Modell S.E. 4200 ist mit einer Anzahl wahlweiser Eigenschaften lieferbar:

Ausgang ±1,4 V bei ±8 mA
±10 V bei 100 mA oder 1 mA

Der Verstärkungsfaktor ist 1 400 bzw. 10 000, d.h. volle Ausgangsleistung bei 1 mV Signaleingabe.

Die Gesamtkonstanz und Genauigkeit ist besser als ±0,1 Prozent des Endwertes bei Abschwächereinstellung von 10 mV. Der Frequenzgang des Verstärkersystems ist besser als ±10 Prozent im Bereich 0...10 kHz. Rauschen ist auf den Eingang bezogen besser als 4 μV_{SS}.

Gegen Zuschlag kann ein Eichsystem eingebaut werden, das sich fernbedienen oder mit einer Datenregistriereinrichtung oder einem Datenverarbeiter verbinden lässt.

Jeder Verstärkermodul ist etwa 51 mm × 165 mm hoch und 330 mm tief; ein 19"-Gestell-Einschub kann bis zu 8 Moduln aufnehmen.

EE 93 767 *für weitere Einzelheiten*

Widerstand-Schaltfeld

Cambion Electronics Products Ltd, Castleton, Nr. Sheffield, Yorkshire

(Abbildung Seite 334)

Das Cambion-"Patch-Ohm" ist ein Miniatur-Widerstandschaltfeld von etwa 32 × 32 mm, auf dem es 56 Standardwiderstandswerte von 51 Ω bis zu 9,5 kΩ mit 5 Prozent Toleranz gibt.

Der "Patch-Ohm"-Widerstand wird nach den modernsten mikroelektronischen Verfahren hergestellt. Der Baustein besteht aus einem Netzwerk hochzuverlässiger Widerstände in Dickfilmform, die von Microtek Electronics (MTE) entwickelte Standardwiderstände mit 5 Prozent Toleranz ergeben. Das Dickfilm-Widerstandnetzwerk wird nach einem neuen Verfahren mit Glas passiviert. Zwischen −80°C und +150°C ist der Temperaturbeiwert des "Patch-Ohm"-Widerstandmaterials 200 × 10^{-6} und die zulässige Belastung für den Dickfilm etwa 2,3 W/cm².

Die verschiedenen Widerstandswerte werden durch unterschiedliches Zusammenschalten der acht Buchsen, die Cambion 1-mm-Durchmesser-"Huckepack"-Schaltkarten aufnehmen, erreicht.

EE 93 768 *für weitere Einzelheiten*

Bauelementtester

Vertrieb: Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 334)

Der Geschäftsbereich Instrumente und Digitalmessungen der Dawe Instruments Ltd kann als Generalvertreter der dänischen Firma A/S Danbridge für Grossbritannien nunmehr den R-C-L-Bauelementtester CPT2 anbieten, den diese Firma herstellt.

Das Instrument ist eine direktanzeigende Messbrücke zum Vergleichen von Impedanzen und Phasenwinkeln über einen breiten Impedanz- und Frequenzbereich. Sondermerkmale des Gerätes sind seine hohe Genauigkeit und Konstanz in Verbindung mit der einfachen Bedienung.

Der RC-Oszillator speist eine Brückenschaltung bei Frequenzen von 100 Hz, 1 kHz, 10 kHz und 100 kHz; direktablesbare Anzeigen erfolgen auf zwei getrennten Messgeräten.

Typische Anwendungsbeispiele sind Abnahme, Fertigung und Klassierung von Bauelementen und Schaltungsteilen, Messen von Spulen, Kondensatoren sowie Elektrolytkondensatoren mit überlagertem Gleichstrom. Das Instrument ist auch für Testen und Justieren mechanisch gekuppelter Bauelemente geeignet.

Im Labor bestehen Einsatzmöglichkeiten beim Messen langfristiger Konstanz, dem Einfluss von Temperatur und Feuchtigkeit, beim Vergleich von Impedanzen mit grosser Genauigkeit und dem Messen von Phasen.

Der Impedanzmessumfang ist 3 Ω ... 5 MΩ im 0,3-V-Bereich und 100 Ω...5 MΩ im 3-V-Bereich. Hochimpedanzwerte können nach Umeichung mit gleicher Genauigkeit bis zu 50 MΩ gemessen werden.

EE 93 769 *für weitere Einzelheiten*

Photorelais

O.P. Instruments Ltd, 246 Sutton Common Road, Sutton, Surrey

(Abbildung Seite 334)

In dieser neuen Baureihe von Photorelais sind alle Baugruppen in wetterfeste Gehäuse aus Aluminiumguss eingebaut. Auswechselbare Druckschaltungs-Einsteckkarten und -relais werden durchgehend verwendet. Die meisten Systeme sind mit Warneinrichtungen ausgerüstet, die bei Ausfall der Lampe oder der Lampenstromversorgung Alarm geben.

Ausser der üblichen Ein-Aus-Arbeitsweise können die verschiedensten Schaltungskarten Verzögerung vor Relaisansprechen oder geregeltes, anhaltendes Schliessen des Relais veranlassen.

Die meisten Baugruppen sind für die normale Licht ein—Licht aus-Arbeitsweise oder Betätigung durch momentane Lichtimpulse lieferbar. Die Schaltungen können auch für Betrieb mit zwei Fotozellen ausgelegt werden; in diesem Fall schliesst Licht auf eine der Zellen das Relais, während Licht auf die andere Zelle es öffnet.

Die verschiedensten Fotozellen- und Lampenköpfe sind lieferbar. Unter günstigen Bedingungen können Miniaturköpfe Gegenstände von nur 1 mm Durchmesser nachweisen.

Die Geräte sind im allgemeinen für Netzbetrieb gebaut; es können aber auch tragbare, batteriebetriebene Systeme geliefert werden.

EE 93 770 *für weitere Einzelheiten*

Höchstspannungs-Konstantstromversorgung

Derritron Instruments Ltd. Parklands, Caincross, Stroud, Gloucestershire

(Abbildung Seite 334)

Derritron Instruments Ltd glaubt, dass die nunmehr angebotene Höchstspannungs-Stromquelle im Gehäuse für ihre Leistung und Qualität die billigste bisher lieferbare ist.

Die Stromversorgung E.0411 gibt zwischen 500 V und 3 kV stufenlos regelbaren Gleichstrom bei bis zu 10 mA ab. Der Regelfaktor ist 2 000 : 1, und die Restwelligkeit ist bei Vollast auf 17 mV_{SS} begrenzt. Die Ausgangsimpedanz ist nicht höher als 100 Ω.

Modell 0411 ist als allgemeine Laborstromversorgung gedacht und an eine 19″-Frontplatte angebaut, und auf Wunsch kann das Gerät für Gestelleinbau aus dem Gehäuse ausgebaut werden. Der Ausgang ist völlig isoliert und auf positive oder negative Spannung umschaltbar.

Die strengsten Schutzmassnahmen wurden getroffen, um sicherzustellen, dass alle Buchsen, Schalter und Potentiometer vom Chassis isoliert sind.

EE 93 771 *für weitere Einzelheiten*

Differentialverstärker mit hoher Verstärkung

Tektronix U.K. Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Abbildung Seite 335)

Durch vereinfachte Gleichstromsymmetrierung, Breitbandempfindlichkeit und vielseitige Frequenzgangwahl ist der Typ 1A7 ein nützlicher Universaleinschub mit hoher Verstärkung für die Tekronix-Oszillografen der Baureihen 530, 540, 550 und 580 (ein Adapter Typ 85 ist erforderlich).

Der Gesamtbereich der geeichten Ablenkempfindlichkeit ist von 10 µV/cm bis zu 10 V/cm mit 0 … 500 kHz Bandbreite für alle Ablenkfaktoren.

Die Verstärkerbandbreite kann durch zwei Frontplattenschalter eingestellt werden, die 8 hochfrequente und 7 niederfrequente 3-dB-Punkte geben.

Eingangsrauschen ist maximal (nicht einschliesslich Drift) 3,3 µV und die Gleichstromdrift bei konstanter Umgebungstemperatur und Netzspannung im Durchschnitt über 10 Stunden nicht grösser als 200 µV/h.

Typ 1A7 kann Gleichtaktsignale mit Amplituden bis zu 200 V_{SS} und Frequenzen bis zu 100 kHz mit einem Diskriminationsfaktor von 50 000 : 1 unterdrücken.

Andere Merkmale sind die Fähigkeit für Differenzverschiebung mit bis zu 300 mV –, eine Differenzübersteuerungslampe und eine Ausgangssignalbuchse.

EE 93 772 *für weitere Einzelheiten*

Schaltungskarten

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 335)

A.P.T. Electronic Industries Ltd kündigt eine Ergänzung ihrer Bauelemente-Auswahl für das modulare MINAR-Chassis-System an.

Die mit C.B.1 bezeichnete Schaltungskarte ist eine phenolharzisolierte, 163 × 108 mm grosse Platte oder Karte, in die ein Muster von Löchern mit Abstand nach dem internationalen Rastergrundmass 2,5 mm gestanzt ist. Aufbauen beliebiger elektronischer Schaltungen auf der Karte erfordert wenig oder gar kein Bohren, da alle Bauelemente mit Anschlussdrähten direkt auf die Karte montiert werden und auf der Rückseite mit den eignen Drähten zusammengeschaltet werden können. Man kann sie auch mit Lektrokit-Lötstiften verdrahten, die in die Löcher gepresst werden und, wo erforderlich, zweiseitige Anschlusspunkte bilden.

Die Schaltungskarte ist für Montage auf zwei Zugstangenträger hinter der Frontplatte der einzelnen Moduln konstruiert. Jeder Modul kann ein oder zwei Karten aufnehmen.

Das modulare Chassisaufbausystem MINAR (MINiature Apparatus Rack) wurde in der Forschungsabteilung der British Broadcasting Corporation für die versuchsmässige Konstruktion von Stereo- und Farbfernsehapparaten entwickelt; es wird von A.P.T. hergestellt und als elegantes, dauerhaftes und vielseitiges System für den versuchsweisen Schaltungsaufbau vertrieben.

EE 93 773 *für weitere Einzelheiten*

Fernseh-Laufzeitleitung

Lexor Electronics Ltd, 25-31 Allesley Road, Coventry

(Abbildung Seite 335)

In Erwartung des kommenden Bedarfs an Spezialbauelementen für Farbfernsehen hat LEXOR eine preiswerte Laufzeitleitung entwickelt, die die Funktion eines Tiefpassfilters ausübt.

Die Anwendung erfolgt als Filter für die Phasenkorrektur des Luminanzkanals in bezug auf den Chrominanzkanal. Die exakten Parameter würden von dem benutzten Farbfernsehsystem und der Schaltungstechnik abhängen.

Der abgebildete Baustein ist typisch und gibt bei einer Impedanz von 1,8 kΩ und Grenzfrequenz von 2,5 MHz mit linearem Phasengang eine Verzögerung von 800 ns. Die Einheit hat einen Durchmesser von etwa 8 mm, ist ungefähr 152 mm lang und kunststoffverkapselt.

Bei Bestellung in ausreichenden Mengen könnte der Preis dieser Bausteine äusserst niedrig sein.

EE 93 774 *für weitere Einzelheiten*

AM-FM-Modulationsmesser

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Abbildung Seite 335)

Ein völlig neuer, volltransistorisierter Mehrzweck-Modulationsmesser wird von Marconi Instruments Ltd angekündigt. Das Modell TF2300 wurde entwickelt, um allen Anforderungen für bewegliche und ortsfeste Stationen im Funknachrichtenverkehr sowie den meisten im Rundfunk und der Telemetrie zu genügen.

Der Trägerfrequenzbereich ist 4 MHz … 1 GHz. Es gibt fünf Hubbereiche von 5 kHz bis zu 500 kHz und Modulationsfrequenzen von 30 Hz … 150 kHz. Für Messungen im Tonfrequenzband kann der Modulationsfrequenzbereich mittels eines Bedienelementes an der Frontplatte auf 15 kHz beschränkt werden.

Zur Bestimmung der AM-Tiefe sind zwei Bereiche vorgesehen: 0…30 Prozent und 0…100 Prozent bei Grundmodulationsfrequenzen zwischen 30 Hz und 15 kHz. Für Klirrfaktormessungen wird der Modulationsfrequenzbereich auf 50 kHz erweitert.

Sowohl für FM- als auch AM-Signale steht ein demoduliertes Ausgangssignal mit einem Pegel von 1 mW an 600 Ω für Anzeigevollausschlag zur Verfügung. Für Messungen von Sendern, die mit Vorverzerrung arbeiten, kann ein 50-µs- oder 75-µs-Entzerrungsfilter in die Ausgangschaltung für das demodulierte Signal geschaltet werden.

Zur Vermeidung von Hilfsoszillatorrauschen kann der Hilfsoszillator kristallgesteuert werden; diese Einrichtung ermöglicht auch schnelles Testen in der Fertigung mit bis zu drei Frequenzen. Man kann auch einen externen Hilfsoszillator anschliessen. Eine eingebaute Kristalleichquelle gestattet Nachjustieren der Hubanzeige.

Das Gerät arbeitet mit den üblichen Netzspannungen oder mit 24 V— an positiver Masse.

EE 93 775 *für weitere Einzelheiten*

Zeitintervallmesser

Airmec Ltd, High Wycombe, Buckinghamshire

(*Abbildung Seite 335*)

Der Zeitintervallmesser 369 ist ein preiswertes und volltransistorisiertes Gerät zum Messen sehr kurzer Zeitinkremente im Bereich 100 ns ... 30 ms.

Das Instrument ist im Grunde eine sehr genaue elektronische Stoppuhr, die in Verbindung mit einem konstanten internen Oszillator zählt und durch positive und/oder negative Schritte oder Impulse betätigt wird. Das Gerät kann für einen beliebigen Triggerpegel zwischen 1 und 10 V eingestellt werden und im Ein- oder Zweileitungsbetrieb arbeiten.

Diese Kombination von Einrichtungen ermöglicht das Messen von Impulsdauer, Impulsabstand, Anstiegs- und Abfallzeiten, sowie die Bestimmung des Zeitintervalls zwischen Impulsen auf zwei verschiedenen Leitungen.

Die Anzeige erfolgt auf einem Messgerät mit 127 mm langer Skala und wird für unbegrenzte Zeit bis zur Rückstellung gehalten.

EE 93 776 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Das elektronische Prinzip des Mekometer II von H. H. Bradsell

Zusammenfassung des Beitrages auf Seite 282-287

Mekometer II ist ein elektronisches Entfernungsmessgerät, das auf die einfachste mögliche Weise Entfernungen bis zu 1,5 km misst.

Ein mit nahezu 500 MHz modulierter Lichtstrahl findet Anwendung. Speziell entwickelte Hohlraumresonatoren bestimmen die Modulationswellenlänge unabhängig vom Einfluss des Brechungsindexes der Luft im Strahlweg auf die Geschwindigkeit des Lichtes.

Eine verbesserte R-C-Brückenschaltung hoher Selektivität von S. Bernstein

Zusammenfassung des Beitrages auf Seite 288-293

Einschalten einer aktiven Stufe mit Verstärkung 1:1 in eine beschriebene R-C-Brückenschaltung erhöht das effektive Q der Schaltung auf einen Wert von 3/2 des mit einer Standard-Wien-Brücke erreichbaren. Eine Verbesserung dieser Art ist besonders in Oszillatoren und selektiven Verstärkerschaltungen bei Hör- und Unterhörfrequenzen vorteilhaft.

Schaltungen für Impulsmodulation mittels variabler Hinterflankensteilheit von O. E. Kruse und J. K. Kincaid

Zusammenfassung des Beitrages auf Seite 294-295

Elektrische Schaltungen für Erzeugung und Demodulation von modulierten Dreieckimpulsreihen werden beschrieben. Die zu übertragende Information wird durch die Steilheit der Hinterflanken in der Impulsreihe übermittelt. Im Demodulationsverfahren werden die Impulse mit variabler Steilheit in Impulse mit variabler Amplitude umgesetzt, die dann verstärkt, gestreckt und gefiltert werden. Das Endausgangssignal ist eine gute Wiedergabe der Originalinformation.

Eine vielseitige Digital-Frequenzquelle für den beweglichen Funkverkehr von A. F. Evers

Zusammenfassung des Beitrages auf Seite 296-303

Dieser Beitrag beschreibt eine abstimmbare Präsizions-Frequenzquelle mit phasenstarrem, durchstimmbaren Oszillator. Das Frequenzregelsystem mit Rückführung enthält digitale Fest- und Regelspannungsteiler und vergleicht einen veränderlichen Ausgangsfrequenzteil des durchstimmbaren Oszillators mit dem Festfrequenzteil des Ausgangs eines hochkonstanten Frequenznormals. Diese synthesierende Digital-Frequenzquelle hat sich für einen breiten Anwendungsbereich im beweglichen Funkverkehr, wo kleine Abmessungen, geringe Stromentnahme und hohe Zuverlässigkeit wichtig sind, als geeignet erwiesen. Bei nur geringem Extraaufwand und nur wenig erhöhtem Verbrauch lässt sich ein äusserst reines Ausgangssignal erreichen.

Einige der bedeutenderen Konstruktionsprobleme werden besprochen und die Eigenschaften einiger Labormodelle und Prototypen beschrieben.

Periodenpunkt-Schalter von R. A. Stevens

Zusammenfassung des Beitrages auf Seite 304-306

Bei der Untersuchung von Einschwingvorgängen in Wechselstromschaltungen ist es oft notwendig, in einem vorgewählten Punkt der Speisewellenform zu schalten und diesen Schaltpunkt genau zu wiederholen. Einrichtungen zur Ausführung dieser Aufgabe sind schon früher gebaut worden, in diesem Falle wurde jedoch beschlossen, eine Ausrüstung ausschliesslich mit Festkörpern zu konstruieren. Eine ziemlich weit verbreitete Phasenverschiebungstechnik, in der ein vorgewählter Pegel auf einer Spannungs-Sägezahnschwingung ein Tor schaltet, findet Anwendung. Das Gerät ist kompakt und tragbar; wegen seiner periodischen Betriebsweise ist es besonders für Laborexperimente von Studenten geeignet.

Ein System für Periodendauermodulation und -demodulation von L. Molyneux

Zusammenfassung des Beitrages auf Seite 307-309

In dem Beitrag wird eine Schaltung für ein Periodendauermodulations- und -demodulationssystem beschrieben und dessen Eigenschaften veranschaulicht. Der Hauptvorteil des Systems liegt beim Vergleich mit der Frequenzmodulation—der es sehr ähnelt—in der Einfachheit des dem Demodulator nachgeschalteten Tiefpassfilters, das in einigen Fällen überflüssig ist.

Transistorisierter Impulstreiber für einen Schrittschaltmotor von W. J. Jirafe

Zusammenfassung des Beitrages auf Seite 316-321

Der Beitrag beschreibt die Entwicklung einer transistorisierten Schaltung, die einen Schrittschaltmotor durch Beaufschlagung des Eingangs mit Impulsen nach rechts oder links weiterschalten kann. Jeder Impuls schaltet den Motor um einen Schritt weiter; die Reihenfolge wird mittels einer Schaltanordnung gesteuert, in der die Schaltfolge durch zwei Flip-flops erzeugt wird, deren Ausgänge nach Verstärkung die beiden Phasen des Motors treiben. Sie werden durch einen gemeinsamen Flip-flop getriggert, der durch Eingangsimpulse über eine Reihenfolgelogik getriggert wird. In einer alternativen Anordnung wird Einsatz eines umkehrbaren binären Zählers vorgeschlagen; die günstigen Eigenschaften der ersten Lösung gegenüber der letzteren werden erläutert. Ausserdem wird die Anordnung für einen Motor mit bifilarer Wicklung angedeutet. Weiterhin wird die Arbeitsweise eines Schrittschaltmotors kurz untersucht und seine Anwendung als Positionsservo besprochen.

Analytischer Entwurf eines Thermistor-Netzwerkes für die Gleichstromstabilisierung von Transistoren von H. Singh

Zusammenfassung des Beitrages auf Seite 322-324

Eine analytische Einstellung zum Entwurf von Thermistor-Netzwerken für die Stabilisierung des Gleichstrom-Arbeitspunktes von Transistoren wird dargestellt. Sie stützt sich auf das Prinzip der Thermistor-Kompensationstechnik, die Stabilisieren des Kollektorstroms durch Einregeln der Basis-Emitterspannung des Transistors bei drei vorgewählten Temperaturwerten gestattet. Durch das beschriebene Verfahren wird die Konstanthaltung des Kollektorgleichstroms im Temperaturbereich 20 . . . 50° wesentlich verbessert.

Dieses Verfahren ist eine Alternative zur Entwicklung der Thermistoreigenschaften mittels grafischer Methoden.