ELECTRONIC ENGINEERING

VOL. 37 No. 446

APRIL 1965

Commentary

WHILE the aviation section of the British electronic industry is in some doubt about its future due to the Government's possible cancellation of certain military aircraft, the computer section is likely to be considerably encouraged by a new four-point plan recently announced by the Government.

The Government is anxious that there should be a rapid increase in the use of computer techniques in industry and commerce and that there should be a flourishing British computer industry which can face up to the competition of the powerful American computer industry.

It is of course essential that this country should have its own independent computer industry if the modernization of industry and commerce is to continue, but the difficulty in the past has been the lack of resources to finance such expansion.

The American computer industry has been receiving enormous support for its research programmes by the placing of large scale military and space research contracts whereas in this country the cost of computer research has had to be borne to a large extent by the companies themselves.

The first step is therefore to set up a computer Advisory Unit within the Ministry of Technology to advise on computer requirements of the whole public sector. In addition, proposals for computers required by Government Departments for non-military purposes will be referred to this Advisory Unit for technical appraisal before permission to purchase is given.

Similar procedure will be applied to computers purchased under Government grants by Universities, Colleges of Advanced Technology and Research Councils.

Advice, based on experience thus acquired by the Unit, will be passed on to computer purchasers in local authorities, the nationalized industries and elsewhere, and the unit will work closely with the computer industry in the planning of new developments.

The next step will be to undertake a full-scale review of the computer requirements of Universities, Colleges of Advanced Technology and Research Councils which is already being studied by the University Grants Committee and the Council for Scientific Policy and this will be followed by a new five year programme, which should produce considerable benefits and provide an important stimulus to the computer industry itself.

Details of the programme are to be announced in the coming months and it is expected that a sum of £2M a year will be set aside to support this programme.

Thirdly, it is proposed to initiate further research into computer techniques and the development of new equipment by industry, universities, Government research establishments and the Post Office. Already the National Research Development Corporation, which is very active in the computer field, has come to an agreement with International Computers and Tabulators Ltd for the joint undertaking of a series of development projects based on the further exploitation of the I.C.T. 1900 Series of computers and a sum of £5M has been set aside by the N.R.D.C. for this purpose.

Looking ahead to the 1970s it is planned to sponsor computer research in partnership with industry under an Advanced Computer Techniques Project which the Ministry of Technology will take over in due course from the old Department of Scientific and Industrial Research and a sum of £1M has been allocated while further developments are being examined.

If the computer industry itself is prepared to contribute towards the cost of widening these programmes it is hoped that within the next two years further contracts will be placed with industry which will involve additional commitments by the Government of a further £½M.

Similar contracts will, it is hoped, be placed with universities within the same period for industry orientated research and this too will cost the Government an additional £½M.

Finally it is proposed to establish a National Computer Programme Centre in conjunction with the industry and computer users and form a library by the purchase of existing programmes as well as the commissioning of new programmes required for use in the British computers.

This will lead ultimately, it is hoped, to the standardization of computer language and so make full use of our scarce resources of programming experts.

A Transfer Function Computer

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

The performance of a linear four-terminal network or control system is defined in all circumstances by its transfer function. A new instrument has been designed which will, by measurement directly across the network or system, compute its transfer function in terms of the complex variable p. Hitherto the accepted method of analysis was based on the measurement of phase and gain or transient response. Use of the transfer function computer (t.f.c.), however, overcomes the need for the considerable work involved in deriving the transfer function from such data.

Provided all the parameters are known, the transfer function of a system can be calculated. The t.f.c. then affords a means of measuring the actual transfer function of the system whence term by term comparison enables individual parameters to be checked. Alternatively, if a system or network is to have a specified transfer function the terms may be set up on the t.f.c. Pre-set controls on the system may then be adjusted according to the specification, or the t.f.c. can be used as a piece of test equipment to accept or reject the system.

(Voir page 280 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 287)

IN general, the differential equation of a four-terminal system may be written as follows:

$$\alpha_0 R + \alpha_1(dR/dt) \ldots \alpha_n(d^nR/dt^n) = \beta_0 E + \beta_1(dE/dt) \ldots \beta_m(d^mE/dt^m) \quad \ldots\ldots (1)$$

where $R(t)$ is the response function with respect to time of the system with an arbitrary excitation $E(t)$ and where all coefficients are real constants. The transfer function of such a system may be defined:

$$F(p) = R(p)/E(p) = \frac{\beta_0 + \beta_1 p + \beta_2 p^2 \ldots \beta_m p^m}{\alpha_0 + \alpha_1 p + \alpha_2 p^2 \ldots \alpha_n p^n} \ldots (2)$$

Hitherto, the technique adopted for examining the characteristic of equation (2) has involved the measurement of phase and gain of the steady state response of the system to a sinusoidal input over a finite frequency band. Given the phase-gain characteristic of the system it is possible to derive equation (2) but generally the work involved is lengthy and tedious.

Arising out of an investigation into the use of differentiating circuits for network equalization by G. G. Gouriet[1], the possibility was suggested of measuring the transfer function of a system by producing successive derivatives of the excitation and response as expressed in equation (1). An investigation of this suggestion lead to the development of the instrument described in this article.

Equation (3) shows the form in which the measured transfer function is obtained. All the coefficients and signs are read directly from potentiometer and switch settings.

$$F(pT) = R(pT)/E(pT) = \frac{\pm b_0 \pm b_1pT \pm b_2p^2T^2 \pm b_3p^3T^3 \pm b_4p^4T^4}{(a_0+a_1pT+a_2p^2T^2+a_3p^3T^3+a_4p^4T^4)(c_0+c_1pT+c_2p^2T^2)} \quad \ldots\ldots\ldots\ldots (3)$$

The factor T makes the coefficient values non-dimensional and also determines the frequency range over which the system is tested.

Description

In describing the instrument it is convenient to refer to the numerator and denominator of equation (2) as functions $f\beta(p)$ and $f\alpha(p)$ respectively. Thus equation (2) can be re-written as follows:

$$R(p)/E(p) = f\beta(p)/f\alpha(p) \quad \ldots\ldots\ldots\ldots (4)$$

In Fig. 1, the driving signal $D(p)$, generated in the instrument, is applied to a series of differentiating circuits of equal time-constants. By means of potentiometers, the amplitudes of the driving signal and each of its successive derivatives are continuously and independently variable. Either of the functions $f\alpha(p)$ or $f\beta(p)$ operating on $D(p)$ can thus be set up by adding together all potentiometer outputs. Using two sets of potentiometers both functions can be set up independently to operate on $D(p)$.

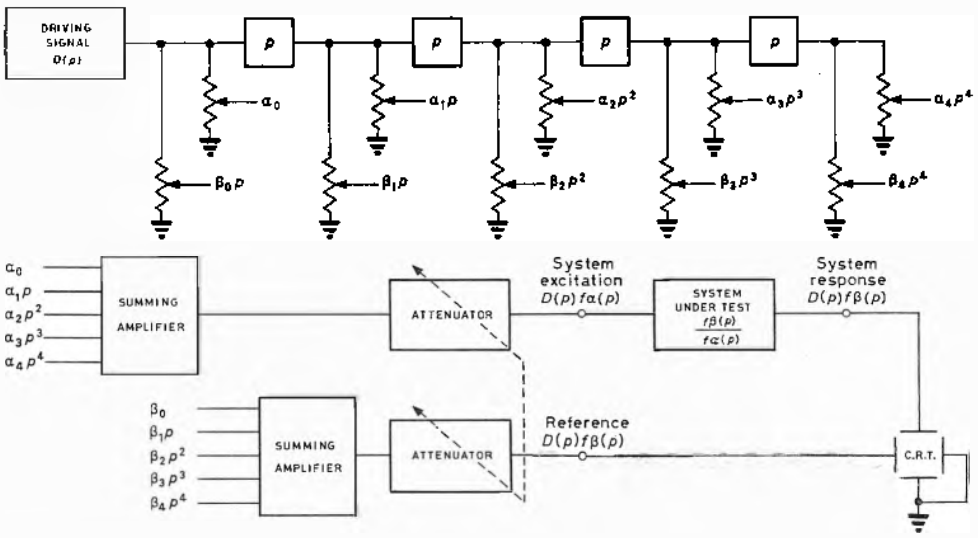


Fig. 1. Principle of operation

The outputs of the two adding circuits give respectively $D(p)f\alpha'(p)$ and $D(p)f\beta'(p)$ of which the former is used to excite the system while the latter serves as a reference signal.

Thus:

$$E(p) = D(p)\,f\alpha'(p)$$

* The Solartron Electronic Group Ltd, formerly Wayne Kerr Laboratories Ltd.

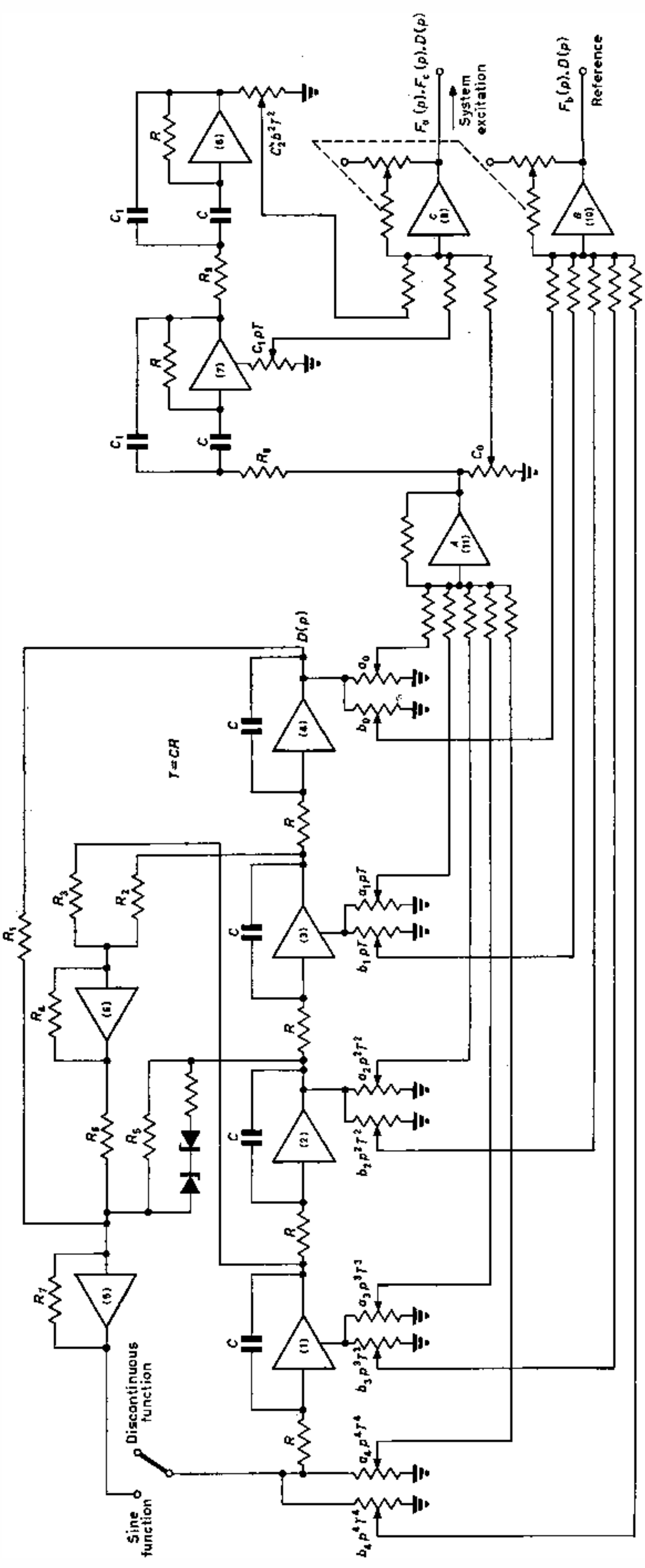


Fig. 2. Arrangement of computer unit

and equation (4) becomes:

$$R(p) = D(p)\, f\alpha'(p)\, \frac{f\beta(p)}{f\alpha(p)} \quad \ldots\ldots\ldots\ldots \quad (5)$$

When all coefficient potentiometers and switches are correctly adjusted $f\alpha'(p) = f\alpha(p)$ and $f\beta'(p) = f\beta(p)$ whence the system response is identical to the reference signal. Coefficient values are read directly off dial indicators associated with the various controls.

To achieve a correct setting, the system response and excitation signals are compared on a cathode-ray tube display over a finite frequency band. For an initial setting, the system response and reference signals are fed respectively to the X and Y deflexion systems and equality is indicated by a straight line at 45°. In all cases it is sufficient to apply consecutively a small number of sine wave signals at suitable frequencies to determine the transfer function.

A more precise setting of the controls is obtained by amplifying the difference between the two signals and displaying the error as a line along the Y axis of the c.r.t. Correct setting is indicated when the line is reduced to a spot.

The computer provides a range of values of T from 0·001 to 4sec in approximately $\frac{1}{3}$ decade steps. Normalized coefficients a, b and c are continuously variable from zero to unity or to 10 according to their position in the equation.

The frequency of sine wave oscillation is determined by the setting of an eleven position switch calibrated ωT from 0·316 to 3·16. The actual frequency of oscillation, dependent also on the setting of T, ranges from 0·015 to 500c/s.

Control of excitation signal level is provided by a two-gang potentiometer which maintains constant the ratio of reference and excitation signal levels.

Driving signals having repetitive ramp or parabolic waveforms may be selected instead of sine waves. Positive or negative step functions may be applied manually to the system for which a three position centre zero switch is provided. Repetition rates of ramp and parabolic functions are determined by the setting of T. (See Figs. 6 and 7). Repetitive step and impulse functions are available as the first and second derivatives a_1 and a_2 respectively of a repetitive ramp function.

The instrument can also be made to display the transient response of a system to a discontinuous function using the repetitive ramp function as a time-base. The system is excited either by a step or an impulse function obtained by raising the appropriate derivative coefficient potentiometer (a_1 or a_2). The transient response appears on the Y axis of the c.r.t. and, since both time-base and excitation functions are derived from the same generator, the display is automatically synchronized.

It is also possible to obtain a phase-gain plot of the system at the fixed frequency settings of the switch ωT. With sinusoidal driving signals,

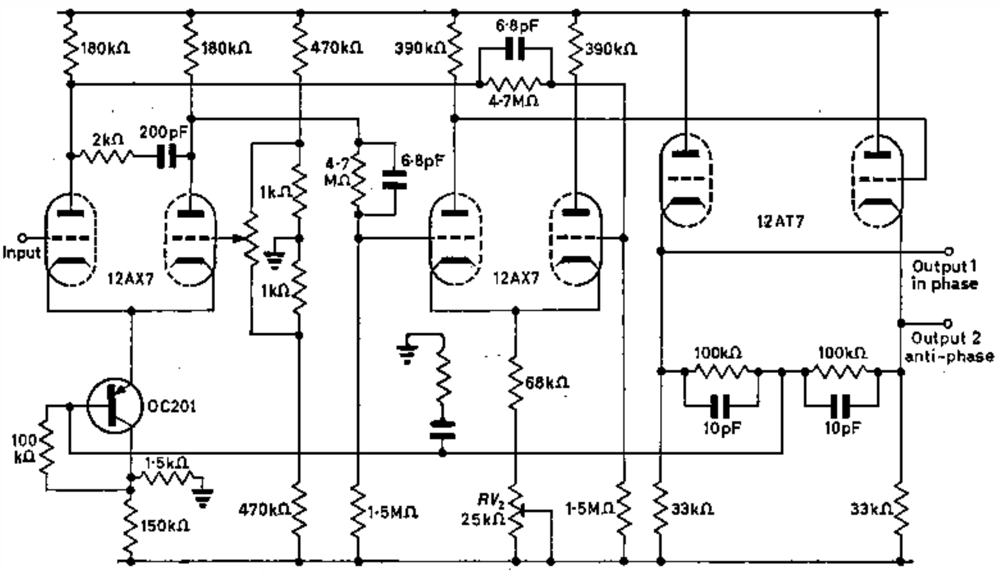


Fig. 3. High gain operational amplifier

each differentiator advances the phase by 90°; thus the coefficient controls vary the amplitudes of a series of phase quadrature vectors. The phase and gain can be derived from a measure of the amplitudes of these vectors which equalize the responses at each frequency.

The instrument thus has three functions; principally to determine the transfer function, but secondly to display transient responses and thirdly to give a measure of gain and phase response.

Circuit Design and Construction

The principal element in the t.f.c. is the computer unit in which the desired functions are produced using high gain d.c. amplifiers. These are plugged in from the rear of the unit while the associated circuits are permanently fitted to the unit chassis. Operational controls are mounted on the 19in front panel.

The power unit provides stabilized 250V positive and negative supplies as well as stabilized heaters at 12·6V for the computer unit. In addition it supplies the e.h.t. voltage as well as heaters and 250V d.c. to the detector unit.

The detector unit contains the c.r.t. with identical X and Y amplifiers with variable sensitivities and shift controls.

Computer Unit

To overcome problems in differentiators, the design has been based on a series of integrators in which the output of the integrator chain may be considered as the input of an equivalent differentiator chain. (See Fig. 2). Two stages of differentiation are, however, used to obtain the terms in the quadratic factor in equation (3).

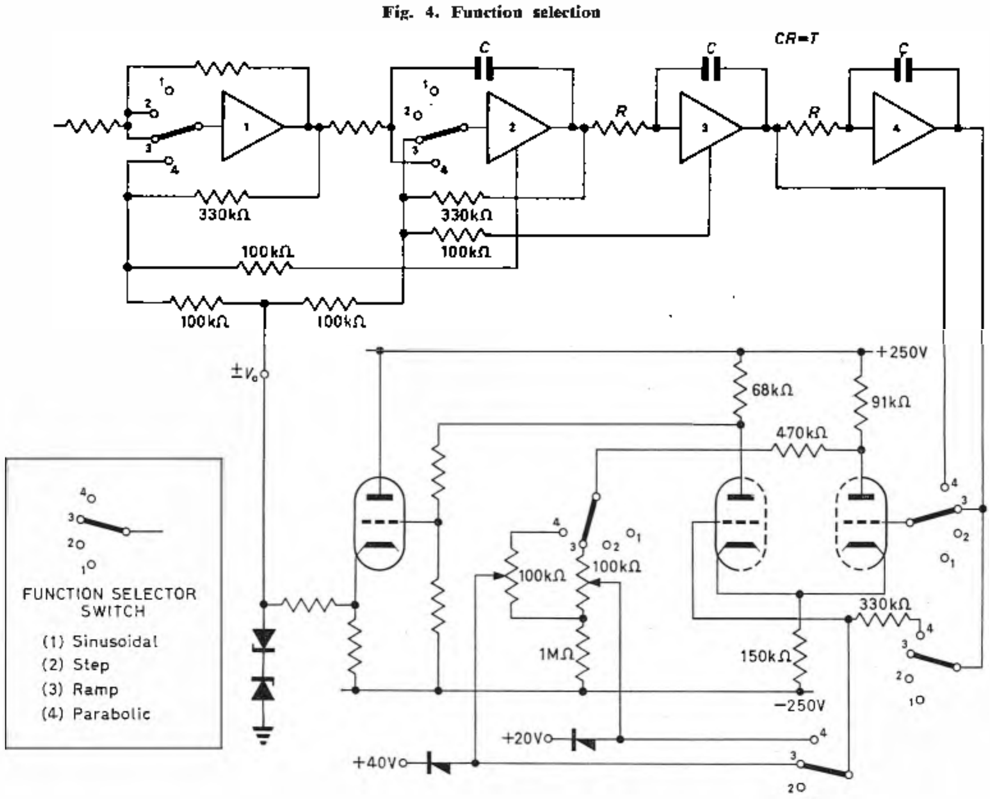
When a sinusoidal driving signal is required, connexion of feedback circuits causes the integrator chain to self oscillate. Feedback proportions are fixed to generate the desired frequency appropriate to the value of T selected.

When ramp or parabolic functions are required the feedback circuits are removed. An impulse is then applied to an appropriate point in the chain so that the required function which is the n^{th} integral of an impulse appears as the driving signal $D(p)$ at the final integrator output.

High gain operational amplifiers are used for integrating, summing and differentiating. Each amplifier has two outputs in phase opposition and may contain a balancing circuit to maintain equal signal amplitudes at the two outputs. (See Fig. 3). Thus the correct phase of signal may be obtained at each integrator output and the facility for sign reversing the numerator terms becomes available.

Coefficient potentiometers are connected to the output of each integrator and to the input of the first integrator.

Fig. 4. Function selection



Two such potentiometers are connected in parallel to each point for numerator and denominator terms respectively. (Reference equation (3)). Each b or numerator potentiometer may, however, be switched to the phase opposed output in order to reverse its sign. Potentiometer outputs are fed to either the a or b summing amplifiers whence the completed series is obtained.

The additional quadratic factor, or c series, in the denominator is provided by a chain of two differentiators which are driven by the signal at the output of the a series summing amplifiers (A_{11}). Outputs from the c series potentiometers together with the a series amplifiers are fed to the total denominator summing amplifier (C_9). Thus the final output, which provides the excitation function for the system $E(pT)$, is the product of the a and c series expressions.

When the function selector switch (Fig. 4) is set to ramp, the second integrator is modified to an amplifier of gain 3·3 and the output of the third integrator is fed back to the amplifier input. A sudden change of d.c. voltage V_o applied to the input of the amplifier causes a change of voltage $V_{(t)}$ at the output of the third integrator of equal amplitude but of longer rise time. The waveform of the voltage at the output of the third integrator is defined by equation (6).

$$V_{(t)} = \pm V_o (1 - e^{-3 \cdot 3t/T}) \text{ Volts} \quad \ldots\ldots\ldots\ldots (6)$$

The waveform at the input of the third integrator is an impulse defined by equation (7).

$$V_{(t)} = \pm 3 \cdot 3 V_o \, e^{-3 \cdot 3t/T} \text{ Volts} \quad \ldots\ldots\ldots\ldots (7)$$

The waveform at the output of the fourth integrator is a ramp defined by equation (8).

$$V_{(t)} = \pm V_o \left[(t/T) + (1/3 \cdot 3)\, e^{-3 \cdot 3t/T} + K_1 \right] \text{ Volts} \quad \ldots\ldots (8)$$

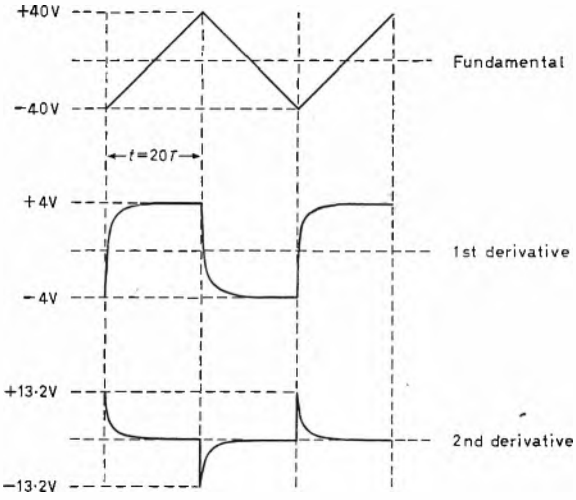


Fig. 5. Waveforms for ramp function
(switch position 3)

When $V_{(t)}$ in equation (8) reaches ± 40V the sign of V_o changes due to the operation of a Schmitt trigger circuit and a new period is started. The Schmitt trigger operates and releases on positive and negative voltages respectively. The resultant waveform defined by equations (6), (7) and (8) is shown in Fig. 5.

For the majority of the period V_t may be considered as changing linearly with time between -40 and $+40$V and the exponential term may be ignored. If $V_o = \pm 4$V then the time t taken for one sweep (i.e. $\frac{1}{2}$ cycle) is $20T$ and $K_1 = 10$.

For a parabolic function (position (4)), the 1st integrator is modified to an amplifier of gain 3·3 and the output of the 2nd integrator is fed back to the amplifier input

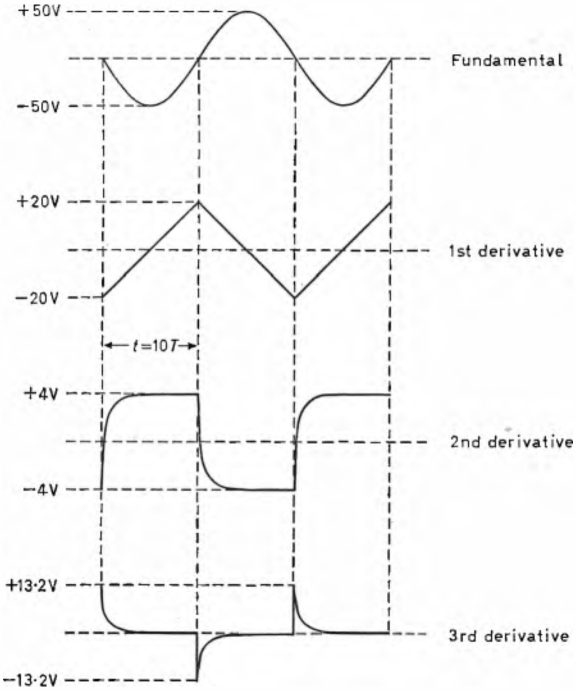


Fig. 6. Waveforms for parabolic function
(switch position 4)

as in Fig. 4. A sudden change of d.c. voltage V_o applied to the amplifier input produces a change at the output of the 2nd integrator $V_{(t)}$ defined by equation (6), and at the output of the 3rd integrator by equation (8).

When $V_{(t)}$ in equation (8) (i.e. at the output of the 3rd integrator) reaches ± 20V the Schmitt trigger operates and reverses the sign of V_o. If V_o is ± 4V then the time taken for one sweep (i.e. $\frac{1}{2}$ cycle) is $10T$ and $K_1 = -5$.

The resultant waveform at the output of the 4th integrator is parabolic and defined by equation (9).

$$V_{(t)} = \pm V_o \left[\tfrac{1}{2} (t/T)^2 - (1/10 \cdot 9)\, e^{-3 \cdot 3t/T} + K_1 (t/T) + K_2 \right] \quad \ldots\ldots\ldots\ldots (9)$$

Ignoring the effects of the small exponential term in equation (9), when $t = 0 = 10T$, $V_{(t)}$ should pass through zero. It is therefore necessary to ensure that any drift voltage K_2 is reduced to zero and this is achieved by feeding the 4th integrator output into the Schmitt trigger in an appropriate manner (see Fig. 4). Waveforms, defined by equation (6), (7), (8) and (9) for the parabolic function are shown in Fig. 6.

Analysis of the Four Integrator Oscillator

The integrator chain in Fig. 2 is shown with the oscillator loop broken at the input to integrator 1 and the open loop transfer function is defined as follows:

$$KG(p) = (R_7 R_4 / R_6 R_3)(1/pT) + (R_7/R_5)(1/p^2T^2) + (R_7 R_4 / R_6 R_2)(1/p^3T^3) + (R_7/R_1)(1/p^4T^4)$$

when the loop is closed at the input to integrator 1, the transfer function becomes:

$$\frac{KG(p)}{1 + KG(p)} = \frac{(R_7/R_1) + (R_7R_4/R_6R_2)\,pT + (R_7/R_5)\,p^2T^2 + (R_7R_4/R_6R_3)\,p^3T^3}{(R_7/R_1) + (R_7R_4/R_6R_2)\,pT + (R_7/R_5)\,p^2T^2 + (R_7R_4/R_6R_3)\,p^3T^3 + p^4T^4} \quad \ldots\ldots\ldots\ldots (10)$$

The denominator of equation (10) has four roots and, for sustained oscillations, two of the roots must be purely imaginary, as follows:

$$p = \pm j\omega$$

Provided all the coefficients of p are positive then the other two roots must have negative real parts. They are most likely to be complex conjugate which implies a damped train of oscillations following closure of the loop. It is desirable that the damping factor of this 'second mode' should exceed 0·4.

The calculation of resistor values for the required frequency ωT is made by substituting $p = j\omega$ into the denominator of equation (10) and equating it to zero as follows:

$$(R_7/R_1) - (R_7/R_5)\,\omega^2T^2 + \omega^4T^4 + j[(R_7R_4/R_6R_2)\,\omega T - (R_7R_4/R_6R_3)\,\omega^3T^3] = 0$$

Real parts: $(R_7/R_1) + \omega^4T^4 = (R_7/R_5)\,\omega^2T^2$ ...... (11)

Imaginary parts: $\omega^2T^2 = (R_3/R_2)$ .... (12)

The second mode roots may now be obtained by dividing the denominator of equation (10) by the known oscillatory roots:

$$p^2T^2 + (R_3/R_2) = 0$$

giving:

$$p^2T^2 + (R_7R_4/R_6R_3)\,pT + ((R_7/R_5) - (R_3/R_2)) = 0 \quad .... (13)$$

which, expressed in the general form becomes:

$$p^2T^2 + 2\zeta\omega_1T\,pT + \omega_1^2T^2 = 0 \quad .... (14)$$

where ω_1 is the 'second mode' frequency.

$$\zeta^2 = (R_7^2R_4^2/R_6^2R_3^2) \times \frac{1}{4\{(R_7/R_5) - (R_3/R_2)\}} \geqslant 0{\cdot}16 \quad .... (15)$$

If R_5 is given the value 10kΩ the values of R_1, R_2, R_3, R_4, R_6 and R_7 may be calculated from equations (11), (12) and (15) for any desired value of ωT. Practical limits are for 0·316 to 3·16.

Control on amplitude of oscillations is effected by means of a limiting circuit associated with R_5. An increase in the value of R_5 causes the imaginary poles in equation (10) to become complex with positive real parts and the amplitude of oscillations builds up. A decrease in the value of R_5 gives these poles negative real parts and oscillations decay to zero.

The value of R_5 has been chosen to cause a rapid build up of oscillations. Two Zener diodes connected in series with a second resistor and in parallel with R_5 effectively reduce the value of R_5 at a given signal amplitude, and any further build up is prevented. This control on signal amplitude at the output of the 2nd integrator is only approximate and a more precise setting is made for each frequency ωT by a slight adjustment to R_7.

Operational Amplifier

Operational amplifiers used for integrating, adding and differentiating have an open loop gain of about 550.

In the integrator chain each amplifier must provide two outputs of opposite sign. This permits the selection of only positive derivatives for denominator terms and either positive or negative derivatives for numerator terms.

Additional common mode rejection is obtained by the use of an external circuit which detects the difference between the two opposite phase outputs. This difference is amplified and fed back to the common cathode of the 1st stage (see Fig. 3).

Monitor Unit

A 2·5in diameter screen c.r.t. with long afterglow is mounted centrally in the monitor unit. X and Y sensitivities are identical and precision attenuators are provided in each channel (see Fig. 7). Neon lamps operate when the spot is off screen and indicate the direction of spot deflexion.

Maximum sensitivity is 30mV/cm in X and Y reducing to 10V/cm in six attenuator steps. Input impedance is 2MΩ for each channel.

Two main inputs to the monitor unit are for the reference and system response signals which are fed to a three-position switch. In position 2, the two signals are fed respectively to the X and Y plates producing a 45° straight line when they are identical. In position 3, the difference between the two signals produces a vertical deflexion reducing to a null on balance, while an X deflexion may be provided by adjustment of the X-expand control. The level of system excitation may be observed by switching to position 1.

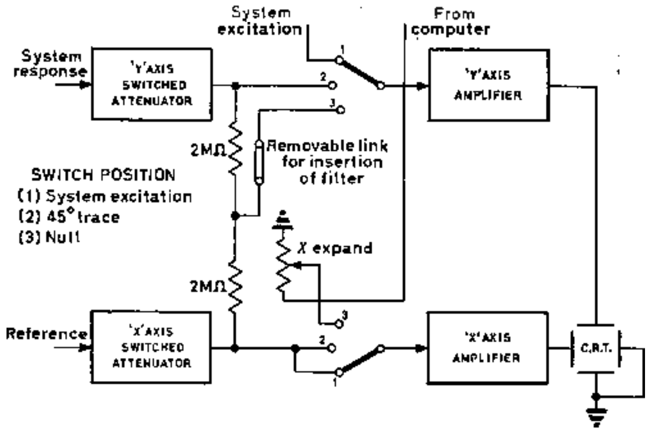


Fig. 7. The monitor unit

When operating the instrument, comparison of the reference and system response signals on the X and Y plates produces an ellipse if the system is linear. Non-linearities cause distortion which varies when the excitation level control is moved. It is often very easy to deduce the source of non-linearity by examining the c.r.t. display.

Operation

Systems with up to 4th Order Denominator

As an example of such a system consider the following transfer function:

$$\text{T.F.} = \frac{1}{1 + 5pT + 10p^2T^2 + 6p^3T^3 + p^4T^4} \quad .... (16)$$

(*a*) *First Stage D.C. Gain*

Initially all coefficient controls are at zero and the a_0 coefficient is then raised to unity. The value of the b_0 coefficient is obtained by balancing the reference voltage with the steady state response of the system to a step signal.

(*b*) *Second Stage* (*First and Second Order Terms Approximately*)

The first and second order terms can be found approximately by equalizing the steady state response at a relatively low frequency e.g. $pT = j\omega T = j0{\cdot}5$.

The system response is compared with the reference signal which in this case is identical to the driving signal, since there are no numerator derivatives. The switch on the monitor unit is set to 45° and the time-constant switch T is adjusted until an ellipse, with its major axis near 45°, appears on the c.r.t. Coefficient controls a_1 and a_2 are then adjusted to produce a straight line at 45° on the c.r.t.

Thus the driving signal is now operated on by $f_a'(j\omega T)$ which must be identical to the denominator of equation (16) at $\omega T = 0{\cdot}5$ in magnitude and phase angle. In this analysis it is assumed that the value of T chosen equals that in equation (16).

The values of the approximated coefficients a_1' and a_2' that would be set up can be found by substituting $pT = j\omega T = j0{\cdot}5$ in the denominator of equation (16) and equating it to $f_a(j\omega T)$ as follows:

$$f_a'(j\omega T) = 1 + j0{\cdot}5a_1' - 0{\cdot}25a_2' = 1 + j2{\cdot}5 - 2{\cdot}5 - j0{\cdot}75 + 0{\cdot}0625$$

$$\therefore 1 - 0{\cdot}25a_2' = -1{\cdot}4375 \qquad \therefore a_2' = 9{\cdot}75$$

$$\therefore j0{\cdot}5a_1' = j1{\cdot}75 \qquad \therefore a_1' = 3{\cdot}5$$

(c) *Third Stage (Third and Fourth Order Terms Approximately)*

The higher order terms can be found approximately using a higher frequency driving signal (e.g. $\omega T = 2$) a_3 and a_4 coefficient controls are adjusted to produce a 45° line and a new function $f_a'(j\omega T)$ results. The values of a_3' and a_4' that would be set up can be found by equating $f_a(j\omega T)$ (which also includes the approximated coefficients a_1' and a_2') to the denominator of equation (13) with $pT = j2$ as follows:

$$f_a'(j\omega T) = 1 + j7 - 39 - j8a_3' + 16a_4' = 1 + j10 - 40 - j48 + 16$$

$$\therefore 16a_4' - 38 = -23 \qquad \therefore a_4' = 0{\cdot}9375$$

$$\therefore j7 - j8a_3' = -j38 \qquad \therefore a_3' = 5{\cdot}625$$

(d) *Fourth Stage (First and Second Order Terms with Greater Accuracy)*

Having found an approximate setting for a_3 and a_4 the first and second order terms can now be found more accurately at the low frequency ($\omega T = 0{\cdot}5$). The values of $a_1' = a_2'$ that would now be set up are follows:

$$a_1' = 4{\cdot}906$$

$$a_2' = 9{\cdot}98$$

which are more accurate than those obtained in the second stage.

(e) This process can be continued indefinitely, but after about the fourth stage 2 per cent accuracy has been reached.

Systems with Third Order Numerator and Denominator

The transfer functions of such systems are of the following form:

$$\text{T.F.} = \frac{b_0 \pm b_1pT \pm b_2p^2T^2 \pm b_3p^3T^3}{a_0 + a_1pT + a_2p^2T^2 + a_3p^3T^3} \qquad \ldots\ldots\ldots\ldots\ldots (17)$$

All derivative terms contained in the numerator may be transferred to the denominator by division. This will in general produce an infinite number of terms some of which will be negative. However, attempts to find the modified transfer function by the method outlined in the previous paragraph will fail when negative terms are encountered. The effect of such terms is immediately apparent when raising a derivative coefficient, making the c.r.t. trace tend away from rather than towards balance.

Coefficient settings may be made in pairs a_1 and a_2 at a low frequency and b_1 and b_2 at a higher frequency. This method is suitable when the system contains a resonance at one frequency and an anti-resonance at a slightly higher frequency. Should these frequencies be interposed then b_1 and b_2 would be set up at the low frequency, followed by a_1 and a_2 at a higher frequency.

The most convenient method for computing the coefficients of such a transfer function uses parabolic functions. Initially, the values of coefficients a_0 and b_0 are determined by application of the step signal as previously described.

If now $D(p)$ is of parabolic waveform (see Fig. 6) the response signal $R(p)$ will have a waveform dependent on the system transfer function. This waveform will contain exponential and possibly sinusoidal components. The effect of raising coefficient controls a_1, a_2 and a_3 to the correct values changes the waveform which then contains only derivatives of the driving signal $D(p)$.

Equation (5) defines the response signal after setting up the a coefficients. For a system having the transfer function in equation (17) $f(\beta)p$ contains zero, first, second and third order derivatives of the parabolic function. Amplitudes

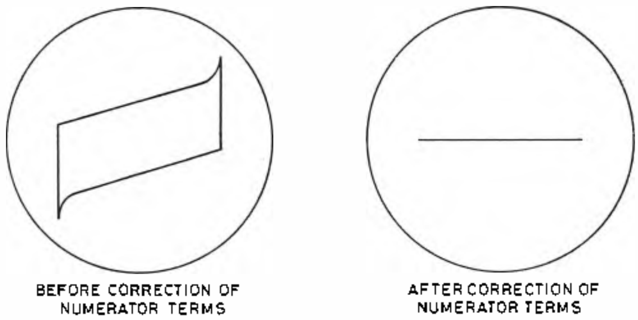


Fig. 8. Display before and after correction of numerator terms

of each derivative are defined respectively, by coefficients b_1, b_2 and b_3.

At this stage, the reference signal will be purely parabolic of amplitude defined by b_0. If the difference between reference and response signals is now examined, it will be seen that the parabolic component in the waveform has been removed. The error signal displayed on the Y axis of the c.r.t. therefore contains only first, second and third derivatives which are respectively ramp, step and impulse waveforms. The effect of applying a ramp signal to the X axis of the c.r.t. is to produce a figure consisting of a parallelogram with two vertical and two sloping sides (see Fig. 8). Superimposed on the top of each vertical side will be a spike, produced by the impulse.

It now remains to reduce this figure to a horizontal straight line. Firstly, an adjustment of b_1 brings the two sloping sides to the horizontal. Secondly, adjustment of b_2 brings these two sides into coincidence. Finally the two spikes at each end of a horizontal line, are removed by the adjustment of b_3.

Thus the operating procedure is firstly to raise coefficients a_1, a_2 and a_3 until such a parallelogram is produced, and secondly to raise b_1, b_2 and b_3 to reduce this parallelogram to a horizontal line.

System Overload

When the error signal exceeds the amplifier saturation level, gain is effectively reduced. Thus, certain coefficients in the transfer function are amplitude dependent. The effect can be observed on the c.r.t. as a change in the displayed figure as signal level is increased. With a low amplitude of sine wave signal a 45° line is produced but this will turn into an ellipse as the level is increased.

If the system under test is a simple second order system, coefficients a_1 and a_2 are inverse functions of the gain parameter. Re-adjustment of a_1 and a_2 will be sufficient to rebalance the display for higher signal levels. A describing function for amplifier saturation may be obtained by examining the values of a_1 and a_2 for given known error amplitudes.

Control Systems with Static, Coulomb and Viscous Friction

The overall friction parameter is given by the relation between motor torque to output velocity. The effect of static and coulomb friction is to make this relationship non-linear particularly at low velocities. In such a system, the first order coefficient a_1 will be output velocity dependent.

The transfer function may be measured using sinusoidal excitation provided that, for each frequency, mean output velocity is kept constant. As signal level is increased, alteration of a_1 value will restore the trace to the 45° line. A plot of a_1 against output velocity will give a describing function for this type of non-linearity.

Noise Rejection

Noise may be attenuated by inserting a low-pass filter into the null detector circuit. Since this operates equally on system response and excitation it has negligible effect on transfer function computation accuracy.

Alternatively, two identical low-pass filters can be inserted into the system response and reference inputs to the detector unit. Such an arrangement would enable the 45° trace to be used.

In many cases, it will be sufficient to construct a low-pass filter having suitable characteristics. The quadratic factors c_1 and c_2 may be used to clear the transfer function of the filter while the remaining coefficients are used to determine the transfer function of the system.

Acknowledgment

The author wishes to express his thanks to the Directors of Wayne Kerr Laboratories Ltd for permission to publish this article.

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A Direct Digital Controller

A new type of direct digital controller, type ARCH 101, has recently been introduced by Elliott-Automation Ltd, and was designed primarily as a replacement for conventional analogue controllers, although additional features such as alarm scanning and communication with other computers have been incorporated.

The standard model is capable of handling up to 30 control loops and may be expanded in modular fashion to nearly 100 loops. It provides proportional or proportional-plus-integral non-interacting control actions. Cascade control facilities and high and low limit alarm scanning are standard features. The programmes for these functions are wired into the machine on plug-in cards which can be replaced by other cards for solving additional control functions.

The action of the programme for a typical machine is quite straightforward and follows the pattern as below:

(a) The input signal from the selected transducer is converted from analogue into digital form.

(b) The measured value is compared with high and low alarm limits. If any point goes into alarm condition, a visual and audible alarm is given and the number of the point in alarm is displayed with a red background.

(c) The measured value is compared with the set value to obtain the deviation.

(d) The deviation is multiplied by the proportional action factor to give the proportional action. The deviation is also separately multiplied by the integral action factor and the integral action is formed. These two actions are summed to give the output control signal.

(e) In cascade control, the output signal of the master control loop is used as a set of the slave control loop.

Input Selection

Each input is connected to the machine through mercury-wetted contact relays. The sequence of selection is determined in advance and is wired-in. The controller accepts electrical signals from any of the conventional types of detecting elements such as thermocouples, resistance thermometers or other transducers giving d.c. voltage or current signals. The selected analogue signal is suitably amplified and converted into a 12-bit binary digital signal.

The machine scans all the points in one second; thus each control loop is up-dated once per second.

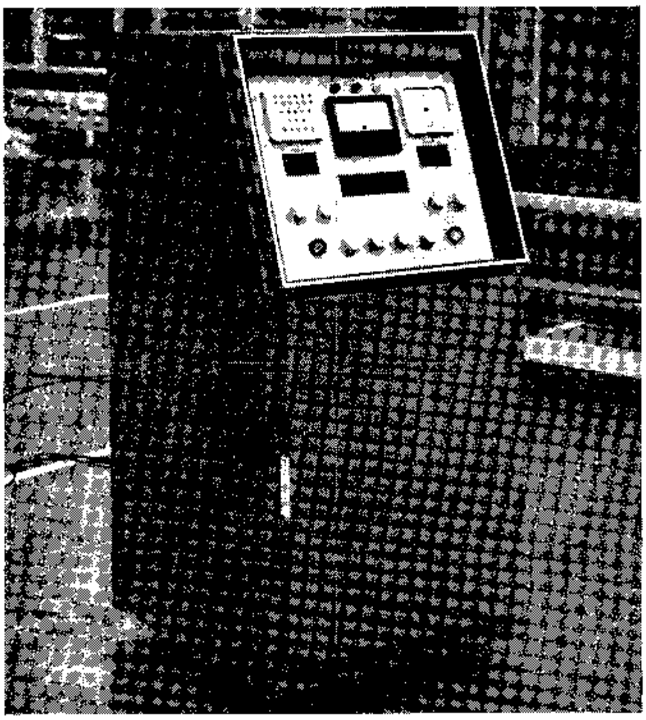
Output Signals

The standard output signal can be either:

(a) In analogue form to suit conventional control valve motor elements, e.g. 1 to 5mA d.c., 4 to 20mA d.c., or 10 to 50mA d.c.

(b) A train of pulses for positioning stepping motor digital transducers.

The Arch 101 Direct Digital Controller



A High Speed Tunnel Diode Decade Scaler

By M. N. S. Swamy*

This article describes in detail the design and operation of a tunnel diode decade scaler, which has wide tolerances for the d.c. biasing currents and the input pulse amplitudes. The main scaler unit itself can operate up to a frequency of about 25Mc/s, whereas the input circuit can handle up to about 10Mc/s. The decade scaler can deliver both positive and negative pulses of 0·5V amplitude.

(Voir page 280 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 287)

ONE of the most useful modes of operation of a tunnel diode is the switching mode. In this mode, it can be made to function as an astable, monostable or bistable device: the resultant circuits can be used in performing memory or logic functions. In the circuits to be developed later, it is in this mode that the tunnel diode is used. Hence, it is of interest to study first the tunnel diode as a switch.

If a tunnel diode is biased by a current source I, then the load line will, in general, intersect the characteristic at three points (Fig. 1(a)). From stability considerations[1-3]

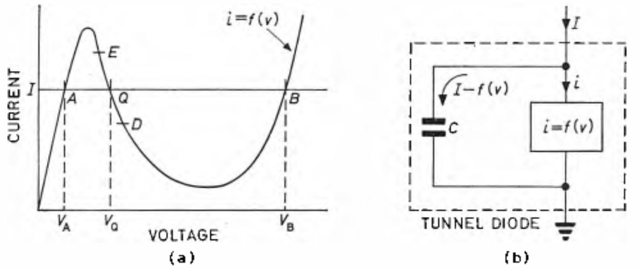


Fig. 1(a). V-I characteristic of a tunnel diode with a horizontal load line on it

(b). Currents through the tunnel diode and the capacitor, under transient conditions

it is known that A and B are stable points, whereas Q is an unstable point. The tunnel diode may be switched from one stable point to the other by the application of a suitable trigger pulse and hence may be used as a bistable device.

Under transient conditions, the current through the tunnel diode capacitor C (Fig. 1(b)) is

$$C(dv/dt) = I - f(v) \quad \ldots\ldots\ldots\ldots\ (1)$$

For any point such as D (between Q and B) $I - f(v)$ is positive and hence is dv/dt. Thus, v increases with time till it reaches the stable point B, where $I = f(v)$. For a point like E (between Q and A), $I - f(v)$ is negative and hence is dv/dt, showing that v decreases with time till it reaches stable point A. If the tunnel diode is at any point beyond B or below A, then it will return to the stable points B or A respectively.

From the above, it is evident that in order to switch the tunnel diode from one stable point to the other, the trigger pulse should take it beyond the 'cross-over point' Q: otherwise it will return to its original state. Hence the minimum charge required to trigger it from A to B is:

$$q = C(v_Q - v_A) \quad \ldots\ldots\ldots\ (2)$$

and that to trigger it from B to A is:

$$q' = C(v_B - v_Q) \quad \ldots\ldots\ldots\ldots\ (3)$$

where C is the effective capacitance across the tunnel diode and is assumed to be independent of v.

Since, in general, $v_B - v_Q$ is greater than $v_Q - v_A$, the charge required to trigger the tunnel diode from the high to the low voltage state is greater than that needed to trigger it vice-versa. Use is made of this charge difference, as will be seen later.

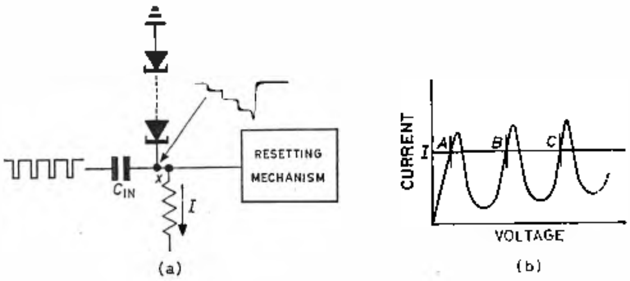


Fig. 2(a). The basic circuit of a tunnel diode scaler

(b). The composite characteristics of the chain of tunnel diodes

The switching time between the two stable states is:

$$t = C \int \frac{dv}{I - f(v)} \quad \ldots\ldots\ldots\ldots\ (4)$$

where the limits of integration are chosen as the 10 per cent and 90 per cent values of the tunnel diode switching voltage.

Principle of Operation of a Scaler

Consider a number of tunnel diodes in series, biased by a constant current source, as shown in Fig. 2(a). The composite characteristic of the chain is shown in Fig. 2(b).

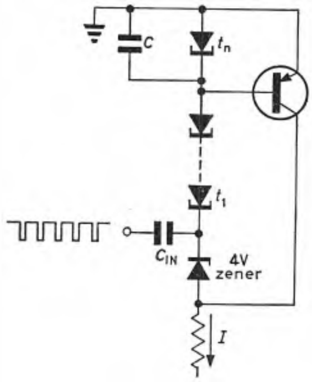


Fig. 3. A tunnel diode scaler using a transistor for the resetting mechanism

The points A, B, C,... are the different stable points. The input is a voltage pulse applied through an input capacitor to the input terminal. This input is differentiated by the input capacitance and the resistance of the tunnel diodes. Thus, a negative and a positive pulse appear corresponding to the negative and positive edges of the input pulse. Corresponding to the negative one, the tunnel diode with the lowest capacitance will switch first, since it needs the

* *University of Saskatchewan, Canada.*

least amount of charge, as given by equation (2). Once the switching action has started, it will inhibit the switching action on the other diodes as there is only a fixed voltage available to be shared among the diodes in the chain. Also, the positive pulse, corresponding to the positive edge of the input pulse, will not necessarily be able to switch this diode back to its low voltage state, since a higher charge is required to do this, as given by equation (3). The effect of the back edge will be dealt with later.

Thus, corresponding to the negative edge of each input pulse, the diodes will flip, one after the other, to their higher states. If a resetting mechanism is arranged which essentially grounds the point X momentarily after all the diodes have switched, then all these will be returned to their original states and the circuit is ready to receive another chain of input pulses. This is the basic principle underlying the design of the tunnel diode scaler[4-6].

Any one particular tunnel diode in the chain may be made to switch last by connecting a small capacitor across it. Hence, conceptually, the easiest resetting mechanism is to trigger a transistor by the switching voltage of this tunnel diode, which has been made to switch last. Then the transistor would remove momentarily the bias current through the diode chain. Then the diodes would switch back, thus turning off the transistor and the chain of diodes would be reset to their initial states. Such a circuit is shown in Fig. 3. But, in this simple configuration, the tunnel diode t_n is loaded by the transistor during switching and hence not permitted to switch completely to its high state.

To reduce this loading while t_n is in the process of switching and to get the necessary delay, a small inductance may be inserted between t_n and the base of transistor. In this case, t_n is allowed to switch completely. Since an inductance is an imperfect delay, the sacrifice of speed is inevitable.

Thus, for reliable and fast operation three conditions must be satisfied. Firstly, every tunnel diode in the chain should be allowed to switch completely. Secondly, when the transistor is conducting, all the tunnel diodes must be returned to the low voltage states. Lastly, the design should ensure that the necessary intervals between the switching of the last tunnel diode and that of the transistor and between the resetting of the tunnel diodes and the turning off of the transistor, are reduced to a minimum. All these conditions may be met by introducing a tunnel diode of a lower peak current between the tunnel diode t_n and the base of the transistor. Such a circuit is shown in Fig. 4, wherein the circuit is made to operate in a sequential fashion, and the following is an explanation of the working of the circuit.

Referring to Fig. 4, $t_1 \ldots t_5$ are all 2mA tunnel diodes, whereas t_6 is a 1mA one. Assume a bias current I_1 of about 1·5mA, a bias current I_2 of about 0·7mA, and let the resistance R be 500Ω. To start with, $t_1 \ldots t_6$ are all in their low voltage states and the voltage across each one of them is about 30mV. Hence, the points M and N are substantially at the same potential and so there is negligible current flowing through R.

The input pulses are applied through the input capacitor C_{in}. Corresponding to the first four pulses, the diodes $t_1 \ldots t_4$ will switch, one after the other, to their higher states as explained earlier. Corresponding to the fifth pulse, the tunnel diode t_5 switches and the voltage at M falls to $-0{\cdot}44$V. As a result, there is now a potential difference of approximately 0·41V between M and N, and a current i of approximately 0·8mA flows from N to M. The current through t_6 has now increased to nearly 1·5mA, which is greater than its peak current, so it switches to its high voltage state, causing a voltage of $-0{\cdot}4$V to appear at N. Since M and N are again almost at the same potential, there is very little current through R and, at the same time, all the diodes are in their high states. The base of the transistor is now at $-0{\cdot}4$V, so that it conducts and completely robs the tunnel diodes, $t_1 \ldots t_5$, of the current I_1. Since there is no bias through these diodes, all of them will be reset. Thus, the point M is at zero potential, whereas N is still at $-0{\cdot}4$V. Hence, a current of 0·8mA flows from M to N, thus switching off t_6 from its high state. This turns off the transistor. The 1·5mA bias current through the diodes $t_1 \ldots t_5$ is restored and all of these diodes return to their normal states. Since M and N are now almost at the same potential, there is no current through R and t_6 is biased to 0·7mA as before. The circuit is now ready for the next train of pulses. At the collector of the transistor, a staircase waveform as shown in Fig. 4 is obtained.

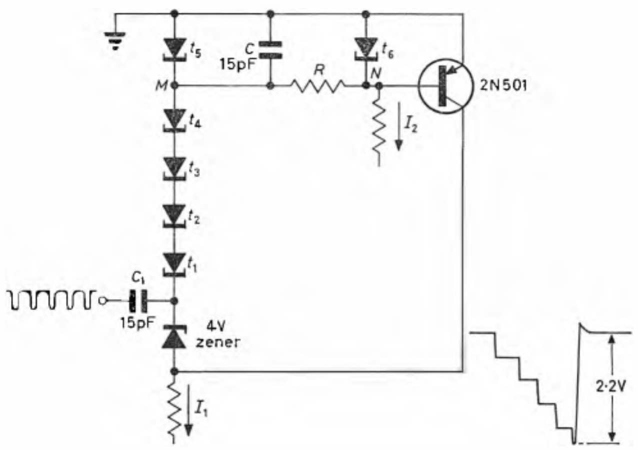


Fig. 4. The circuit diagram of the high speed tunnel diode scaler of five
t_1 to t_5 are 2mA germanium tunnel diodes (1N2969); t_6 is a 1mA germanium tunnel diode (1N1939)

The speed of operation of this scaler is mainly dependent on the resetting time, which includes the switching time of t_6 from its low to high state, the turn-on time of the transistor, the switching time of t_6 from the high to the low state and the turn-off time of the transistor.

The resistance R was chosen experimentally for the best d.c. conditions, so that maximum tolerance was achieved for both I_1 and I_2. This value was found to be 330Ω. For this value of R, I_1 could be varied from 1·1 to 1·7mA, and I_2 from 0·5 to 0·7mA. The scaler works satisfactorily for input pulse amplitudes ranging from 0·5 to 1·1V, for pulses of rise time 3nsec. The circuit could handle pulses whose width was greater than 10nsec. The double pulse resolution for switching between intermediate states was found to be 12nsec, but about 30nsec must be allowed for resetting operation to be completed. Hence the maximum speed of operation of the scaler is about 25Mc/s.

Input Pulse Requirements

It was pointed out in the previous paragraph that the scaler of Fig. 4 works satisfactorily, provided the input pulse amplitude lies within certain limits. The lower limit is the minimum input pulse amplitude that is required to take one tunnel diode from the low to the high state. The upper limit is brought about by the fact that, above a certain amplitude of the input, two of the diodes would be capable of switching at the same time. It was found

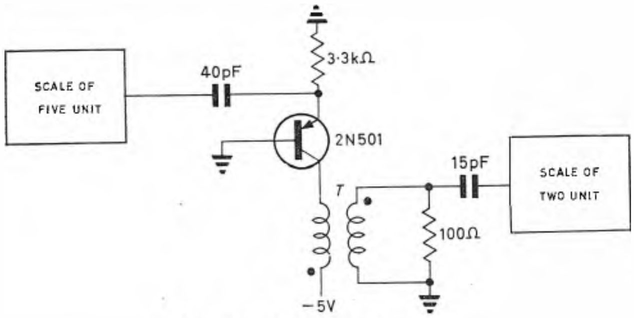


Fig. 5. The decade scaler unit

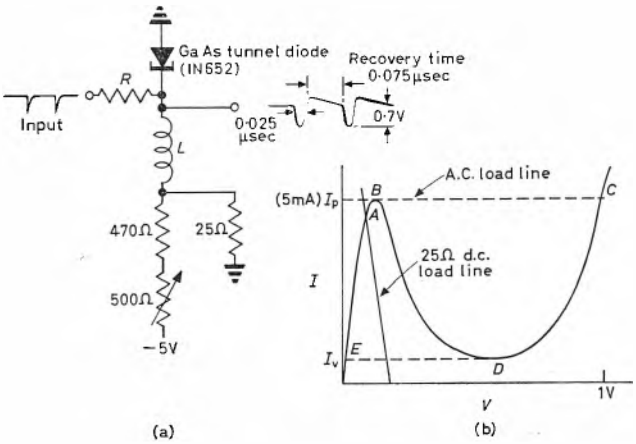
from experiments that, at any given slope of the triggering edge of the input pulse, the range of the input pulse decreased, as the number of diodes in the chain of Fig. 4 was increased. It was also observed that with the increase of the number of diodes in the chain, the tolerance for the d.c. biasing currents decreased. It was seen that if there were eight diodes in the chain, I_1 had to be 1·45mA, I_2 had to be 0·64mA and also, the input pulse amplitude could not be varied very much.

It was also observed from experiments that at any slope of the back edge of the input pulse, the amplitude of the pulse necessary to bring back a tunnel diode in the high state to the low state was greater than that required for the input pulse to take two tunnel diodes simultaneously to their high states. Hence, the effect of the back edge of the input pulse may be ignored in the design considerations of the scaler.

Decade Scaler Unit

From the above considerations, it may be seen that with the configuration of Fig. 4, a decade scaler cannot be designed using ten tunnel diodes in a single chain, if large tolerance for the d.c. biasing currents and the input pulse amplitude is required. Hence, to design a decade scaler, it is necessary to build a scale of 5 and follow it with a scale of 2, or vice-versa. The minimum pulse amplitude necessary for a satisfactory operation of the scale of 5 is more than that needed for the scale of 2. Also, the collector (Fig. 4) swings by about 2·2V for a scale of 5 whereas it changes by 0·9V for a scale of 2. Hence, it is easier to trigger a chain of 2 from a chain of 5 than vice-versa. In the decade scaler described here, a scale of 5 is followed by a scale of 2.

Fig. 6(a). Tunnel diode monostable multivibrator circuit
(b). V-I characteristic of a gallium arsenide tunnel diode with a 25Ω load line on it



It is already seen that at the collector (Fig. 4), a staircase waveform appears. If this staircase is differentiated, then corresponding to each of the first four steps, there will be four small pips. Corresponding to the last one, there will be a small negative pip and a large positive pip. The negative pips can be removed and the positive pulse can be reversed in polarity and used to trigger the scale of 2. This is achieved by the coupling circuit of Fig. 5, where the decade scaler unit is shown.

The staircase output from the scale of 5 is fed to the emitter of a 2N501 transistor through a capacitor of 40pF. The emitter to base diode resistance along with this capacitor differentiates the staircase and the negative pulses are removed automatically at the collector. Corresponding to the resetting positive edge of the staircase, a 15nsec wide pulse of 0·6V amplitude appears at the primary of the transformer. Across the secondary, the same pulse appears, reversed in polarity. A 100Ω resistor is connected across the secondary to obtain the appropriate source impedance to trigger the next stage.

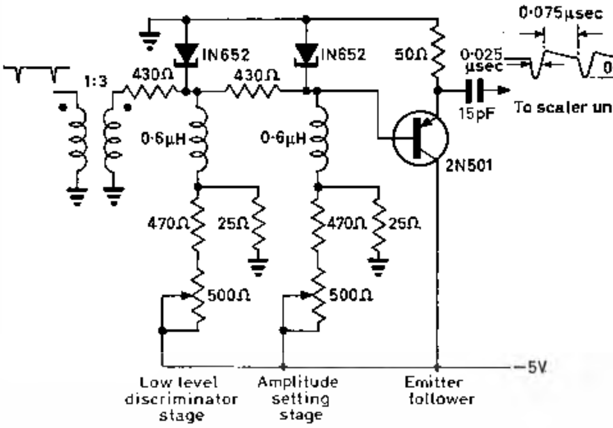


Fig. 7. The complete input circuit

This negative going pulse is then applied through a 15pF capacitor to the scale of 2, which is identical to that of 5, except that 2 tunnel diodes are used in the chain instead of 5. The output at the collector in the scale of 2 unit is fed to the output circuit (Fig. 8), which can supply both positive and negative pulses of 0·5V amplitude and 0·1μsec width.

The Input Circuit

As explained, the scaler can accept a pulse whose amplitude ranges from 0·5 to 1·1V and whose width is greater than 10nsec. Since the pulses to be counted may have varying amplitudes and widths, it is necessary to standardize them ahead of the input to the scaler unit. The input circuit receives the incoming pulses and delivers standard pulses of amplitude 0·7V and width 25nsec.

The basic circuit which will be used to achieve low level discrimination and pulse shaping, or to set the level of the standard pulse is a tunnel monostable multivibrator[6-8], shown in Fig. 6(a). Fig. 6(b) shows the voltage-current characteristic of a 5mA gallium arsenide tunnel diode together with a 25Ω load line. A gallium arsenide diode was used instead of a germanium one, as the switching (jump) voltage of the former is greater than that of the later. The tunnel diode may be switched by the application of a negative trigger pulse, which takes the diode current momentarily to the point B. Then, the diode switches to the point C, as shown by the a.c. load line. Now, the current in the inductance begins to decrease

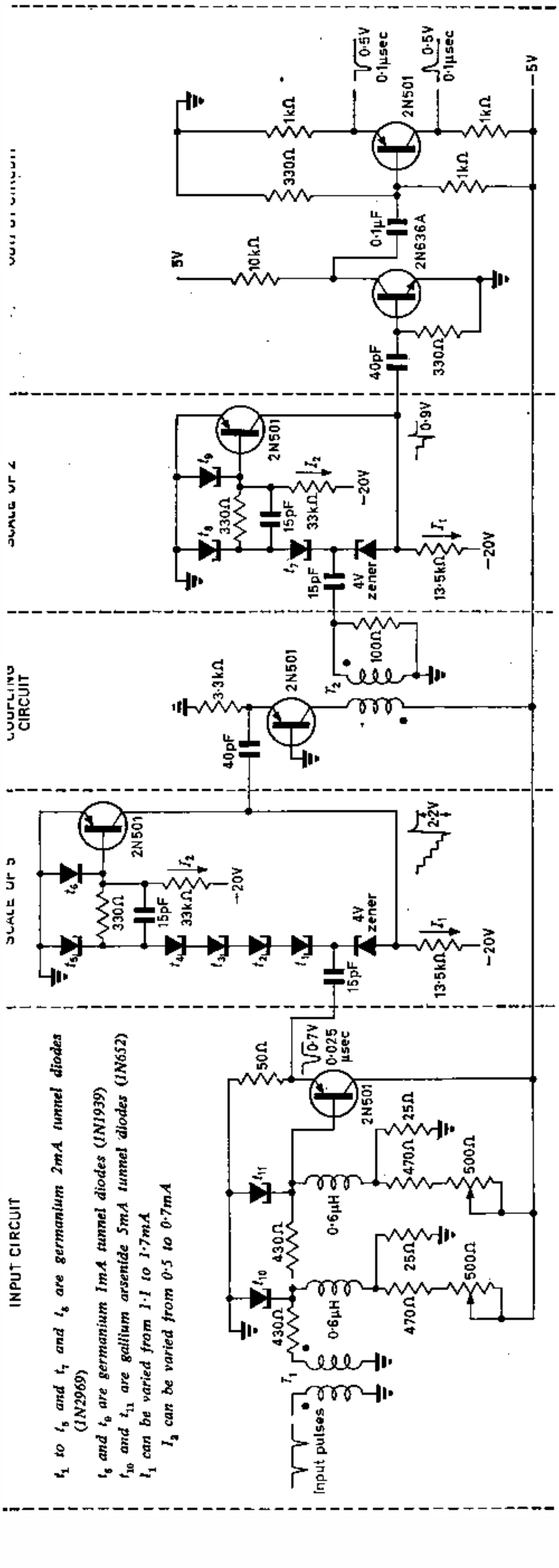


Fig. 8. The complete circuit showing the decade scaler with the input and output circuits

exponentially until it reaches the valley current whereupon the diode switches to the low state. It was found from experiments that the minimum inductance that could be used to get at least 0·7V switching was 0·6μH. For this value of inductance, the pulse width was about 25nsec and the 'recovery time' of the monostable circuit about 75nsec, where the recovery time of the circuit has been defined as the time separation between a pair of pulses that are just resolved when the second trigger pulse has an amplitude 10 per cent above the threshold.

Thus, the speed of operation of this monostable circuit is about 10Mc/s. By using a smaller inductance in conjunction with a faster tunnel diode like the RCA 1N3138, which is believed to have a switching time of less than 1nsec, the speed of the circuit may be increased, so that this circuit could operate at the same speed as that of the scaler unit, which is about 25Mc/s.

The complete input circuit unit consists of a discriminator, followed by a level setting stage and an emitter-follower (Fig. 7). The discriminator stage determines the lower threshold, which may be varied by moving the operating point A up or down by means of the 500Ω potentiometer. The maximum sensitivity which can be realized is 75mV on the 50Ω input side of the transformer. Even though the tunnel diode is a low impedance device, the step-up transformer with the 430Ω resistor is used to give a satisfactory match to the 50Ω line and to attenuate the monostable multivibrator pulse which returns to the previous stage. The use of the transformer enables the circuit to be used with either polarity pulse, while the threshold can be set-up to about 1·2V by means of the 500Ω potentiometer.

The output voltage of an ideal discriminator should be completely independent of the overdrive at the input. The discriminator described here is not ideal in this respect. The output changes from 0·7 to 1·5V, when the input is changed from 75mV to 2V. Hence, the pulse is coupled to another identical stage, namely, the level setting stage. After this stage, the output has a constant amplitude of 0·7V. The emitter-follower is used for the purposes of isolation between the input and the scaler units.

The Complete High Speed Scaler

Fig. 8 shows the complete circuit diagram of the decade scaler including the input and the output circuits. By connecting a small capacitor (of about 15pF) across the resistances which appear between t_5 and t_6 and again, between t_8 and t_9, the changes across t_5 and t_8 may be made to appear immediately across t_6 and t_9 respectively, thus increasing the speed of the scaler. At the same time these capacitors in series with the capacitances of the diodes t_6 and t_9 will ensure that t_5 and t_8 switch last. Hence, the capacitors which were previously connected across t_5 and t_8 (Fig. 4) may be removed. This change is incorporated in Fig. 8.

Conclusions

The decade scaler unit described here has a resolving time of about 40nsec. It has wide tolerances for the d.c. biasing currents and for the input pulse amplitude. It can work satisfactorily for input pulses of width greater than 10nsec and amplitudes ranging from 0·5 to 1·1V. The circuit is relatively simple and, since it employs tunnel diodes, provides simplicity, low power consumption, high speed and lends itself to miniaturization.

The input circuit is, in essence, a tunnel diode monostable multivibrator. The lower level discriminataion of this circuit is about 75mV and the resolution 0·1μsec. The input circuit delivers standard pulses of width 25nsec and amplitude 0·7V. It is expected that the speed of the input circuit could be improved by using faster tunnel diodes. The output circuit can supply both positive and negative pulses of amplitude 0·5V and width 0·1μsec.

The decade scaler may be used as a pre-unit to existing fast scalers or may be followed by a similar decade scaler unit. The principle of the monostable circuit could be made use of in developing coincidence and anti-coincidence circuits. The tunnel diode chain principle may also be made use of in designing an analogue-decimal convertor.

Acknowledgments

The author wishes to express his gratitude to Professor N. F. Moody for his guidance, encouragement and criticisms offered throughout the course of this work. Thanks are also due to the National Research Council of Canada and the University of Saskatchewan for funds and facilities made available to the author during the course of the work.

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Daylight Radar Displays

The Ministry of Aviation has ordered eighteen of the new Marconi high-brightness radar displays, which can be used under any lighting condition, including direct, bright sunlight. This is a major step towards reducing the pressure on air traffic controllers, particularly at the busier international airports.

This display system is over 1 000 times brighter than a conventional radar display and at least 20 times brighter than systems using television scan conversion techniques. The 18 displays on order have a 5in display tube, and will be used as 'Distance From Threshold Indicators'. They will show a radar picture of the aircraft approach path to touchdown, and will be situated in the brightly lit 'glasshouse' at the top of the control tower where the final approach is monitored by the airfield controller. Previously, this has been impossible, without the use of cumbersome tube masks or filters.

This order was received after extensive evaluation at both London Gatwick and London Heathrow airports. The prototype has been in 24h operation at London Heathrow for the past twenty months and has proved to be a most useful aid, particularly for parallel runway operation.

It is an essential feature of a.t.c. radar displays that the traces which appear on the tube face should remain fairly bright for at least the duration of a single scan, and preferably for considerably longer. This effect shows successive positions of moving aircraft and produces a 'tail' on each trace giving an indication of the speed and direction of each aircraft. Unfortunately phosphors which give long persistence do not produce large light outputs and have rather short lives. The Marconi Distance From Threshold Indicator solves this problem by using a special direct view storage display tube. This tube combines the effect of long persistence with very high brightness and is clearly visible in all lighting conditions, including bright sunlight.

Because the persistence of this tube is not derived from the phosphor, the tube has a very long life. The original tube, installed at London Heathrow in March 1963, was replaced a year later after 8 500 hours use. The second tube has now been in continuous operation for approximately 6 500 hours. Normal radar display tubes seldom have a life of more than 1 500 hours.

Active development is proceeding on a similar radar display for use in Approach Control and Area Control rooms. This display will have a much larger storage tube than the D.F.T.I. and will dramatically change the environment in which air traffic controllers are at present working. During the day the curtains may be opened and at night plenty of artificial light can be provided to eliminate the constant strain of working in perpetual semi-darkness or with cumbersome tube masks or filters.

The Direct View Storage Tube

The Distance From Threshold Indicator is based on the use of a direct view storage tube (d.v.s.t.) which produces a very bright radar picture using a high-efficiency short persistence phosphor but which contains a storage element to introduce persistence to the picture.

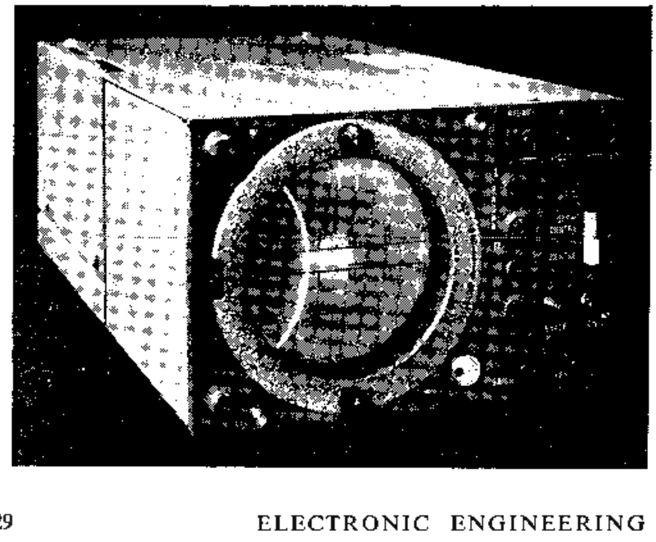
The tube used in the Marconi bright display is an English Electric type E 702A direct view storage tube having a screen luminance of over 1 000ft lamberts under normal conditions and up to 2 150ft lamberts with increased final anode voltage. This tube contains both the viewing surface and the information storage element in a single glass envelope. The viewing surface consists of a normal high-efficiency short persistence phosphor on the inside of the glass faceplate, while the storage element consists basically of an extremely fine metal mesh placed immediately behind the phosphor surface and with a layer of dielectric material deposited on the side furthest from the phosphor.

Two electron guns are mounted in the neck of the tube. One, known as the 'flood' gun, directs a low velocity stream of electrons through electrodes which collimate them to arrive at the storage mesh in the form of a parallel beam normal to the mesh and covering its entire surface area uniformly. These electrons have a sufficiently low velocity to produce a secondary emission of electrons from the storage mesh which is smaller than the beam current, and so tends to drive the mesh negative, preventing the 'flood' beam electrons from penetrating the mesh and reaching the phosphor.

The other gun, the 'writing' gun, produces a beam of high velocity electrons, focused to a fine spot at the storage mesh. This beam passes through a deflexion system and can be modulated as in a normal cathode-ray tube. When this beam strikes the storage mesh, the secondary emission is high and a positive charge is left on the mesh, allowing 'flood' beam electrons to pass through and strike the phosphor.

The application of a positive pulse to the storage mesh will erase the image traced out by the 'writing' gun, and hence the picture on the faceplate. Controlled pulses can be used to cause the picture to decay slowly and the effective persistence of the tube is, therefore, variable up to a maximum of ten to fifteen minutes.

The high brightness radar display



A TRANSISTORIZED INTEGRATOR

By J. A. Effenberger*, D.Ing., A.M.I.E.E., and R. Steele†, B.Sc., A.M.I.E.E.

A gated integrator producing positive or negative ramp voltage is considered. A practical all transistor circuit consisting of a direct coupled amplifier connected in a normal resistance and capacitance feedback network with a gating electronic switch for discharging the capacitance is presented and discussed. During the flyback period the maximum of 28V peak output voltage is discharged to 5 per cent after approximately 50µsec and thence to less than 10mV after further 60µsec. The linearity of the ramp output voltage is better than 2 × 10⁻⁴.

(Voir page 280 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 287)

FIG. 1 shows the block diagram of the integrator in which

A_1 represents the integrating amplifier,

A_2 represents the reset limiting amplifier,

S_1 represents an 'off-on' gating switch (planar transistor),

S_2 represents an 'on-off' gating switch (alloy transistor),

A_3 represents the gate amplifier.

Amplifier A_1 should satisfy the following requirements:

(a) It must have a high input impedance and should have a low output impedance.

(b) It must have a satisfactorily low drift over the required temperature range.

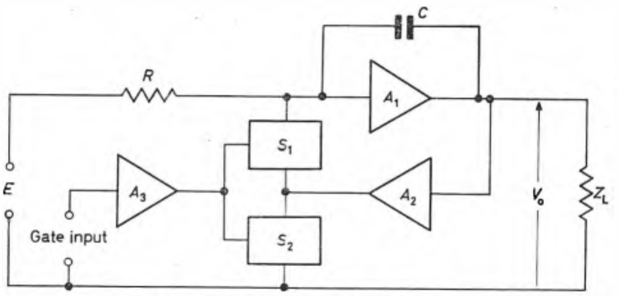


Fig. 1. Arrangement of integrator

(c) Its overall voltage amplification must be high and better than 74dB if the linearity of the integrator voltage output is to be better than 2×10^{-4}. Input and output must have opposite polarities.

(d) The noise generated must be low and preferably less than 3mV at the integrator output ($\simeq$ 0·01 per cent of V_{out}).

(e) Its bandwidth is related to the required accuracy of the initial portion of the ramp and to the p.r.f. of the gate. Generally it need not be very wide.

Amplifier A_2 must, (a) cope with relatively high input voltages of either polarity, and limit with small input voltages, (b) be non-phase reversing, (c) not appreciably load the main amplifier.

Amplifier A_3 is a pulse amplifier which controls the electronic switches S_1 and S_2.

During the integration period S_2 is closed, clamping the output from A_2 to earth while S_1 is open, enabling the input of A_1 to act as a 'virtual earth'. During the 'flyback' period S_2 is open, but S_1 is closed and the capacitor can discharge through A_2.

* *Cossor Electronics Ltd.*

† *The Marconi Co. Ltd.*

Theoretical Considerations

Scanning Period

In a perfect integrator the whole input current would charge the integrator capacitor C. In this integrator errors are introduced due to currents flowing into the reset system and into the main amplifier. It is shown in Appendix (1) that the ratio of actual output voltage to the ideal output voltage is:

$$(V/V_{op}) \simeq \left[1 - \frac{R_{2N} R E_L}{R_{20} R_{1F} E}\right]\left[1 - \frac{t}{2CrG_1}\right] \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

where $r \simeq \dfrac{RR_i}{R + R_i}$

By making the reset amplifier limit early in the initial part of the scan the error current flowing into the reset system is nearly constant and very small due to the very

Symbols

E = Integrator input e.m.f.

E_L = Limited output voltage of the reset amplifier

e_{D1} = d.c. drift e.m.f. of A_1 amplifier referred to its input

e_{D2} = d.c. drift e.m.f. of A_2 amplifier referred to its input

V_o = Actual integrator output voltage

V_{op} = Ideal integrator output voltage

V_{om} = Integrator output voltage due to e_{D1} and e_{D2} during the flyback period

V_{od} = Integrator output voltage during the scan period due only to e_{D1} and e_{D2}

R = Integrator input resistance

R_i = Input resistance of A_1 amplifier

R_{1F} = Off resistance of S_1 switch

R_{1N} = On resistance of S_1 switch

R_{2F} = Off resistance of S_2 switch

R_{2N} = On resistance of S_2 switch

R_{20} = Output resistance of A_2 reset amplifier when limiting

R_{20}' = Output resistance of A_2 reset amplifier when not limiting

C = Integrator capacitor

G_1 = Voltage gain of main amplifier A_1

G_{10} = d.c. voltage gain of A_1

G_2 = Voltage gain of reset amplifier A_2

G_{20} = d.c. voltage gain of A_2

τ = Time from the beginning of flyback period to the time just before A_2 ceases to limit

V_{OD} = Integrator output voltage at the end of the flyback period due to e_{D1} and e_{D2}

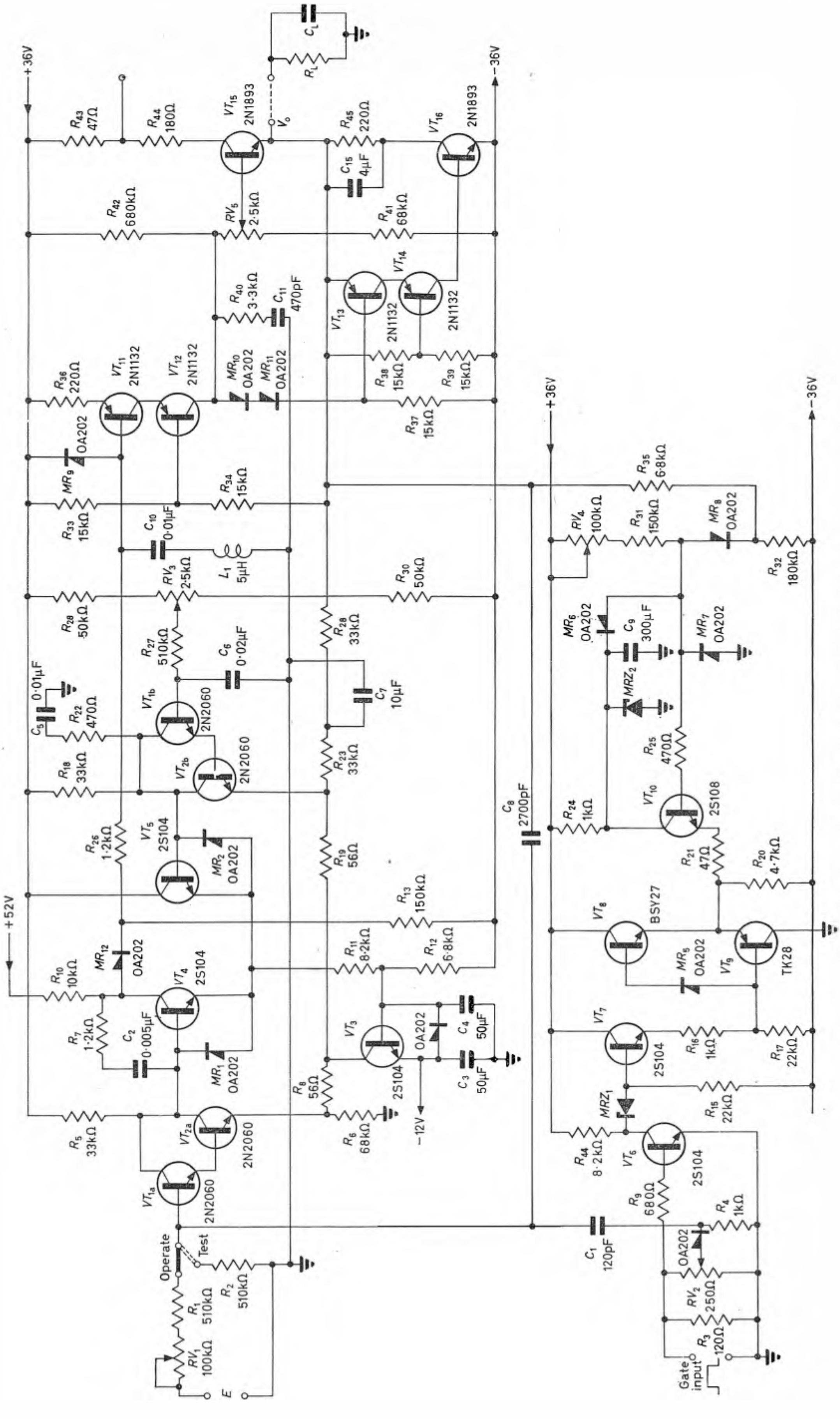


Fig. 2. Circuit of complete integrator

high off-impedance of the planar transistor used as switch S_1, and the very low on-impedance of the allow transistor functioning as switch S_2. However, the second term in the above equation is the most influential as the main cause of the error is due to the amplifier A_1.

FLYBACK PERIOD

It is shown in Appendix (2) that while the reset amplifier is limiting, the capacitor is being discharged by approximately constant current for a time.

$$\tau = C\,[R_{20} + R_{IN}]\,E/E_L \qquad (2)$$

If $G_2 = 1$ then τ is the time for the integrator output voltage to fall by $[1 - (E_L/G_2E)]$. When the integrator capacitor has been totally discharged, the output voltage has been reduced to:

$$V_{om} = \frac{E\,(R_{20}' + R_{IN})}{R_{20}R} \qquad (3)$$

which is the value from which the next scan will start instead of zero as in the ideal case.

EFFECT OF AMPLIFIER DRIFT E.M.F.'S ON INTEGRATOR PERFORMANCE

If the effects of drift, caused by temperature variation, can be considered as generators of e.m.f.'s e_{D1} and e_{D2} then, as shown in Appendix (2), the errors in the integrator output voltage, contributed only by the amplifiers' drift e.m.f.'s are:

(a) At the end of the flyback period.

$$V_{OD} = \mp e_{D2} \pm (e_{D1}/G_{20}) \qquad (4)$$

(b) During the scan period.

$$V_{od} = V_{OD} + (e_{D1}t/RC) \qquad (5)$$

V_{OD} can be reduced by increasing the gain of the A_2 amplifier G_{20}, but this gain cannot compensate for the slope of V_{od}, namely:

$$e_{D1}/RC$$

Design of the Main Amplifier

Fig. 2 shows the detailed circuit of the integrator. To provide a satisfactory high input impedance it is necessary to use emitter-followers, in the first stage of the amplifier. Similarly to achieve low output impedance it is necessary to add a complementary emitter-follower to the amplifier. This amplifier is a variant of a circuit published by Middlebrook and Taylor[1]. It consists of a differential stage $VT_{2a,b}$, with a common emitter current regulator VT_3 controlled by a part of the mean voltage of the emitters of VT_4 and VT_5 in the second stage. This stage converts the differential signal into a single-ended signal available from the collector of VT_4. Cascode-connected VT_{11} and VT_{12} provide the voltages required by the compound output emitter-follower consisting of $VT_{13}, VT_{14}, VT_{15}$ and VT_{16} normally operating in class-A conditions. Because the output voltage is required to be of the opposite polarity to the input voltage overall feedback could not be used, as in the Slaughter[2] circuit, between the output and the other base input and had to be returned to the emitter of VT_{2b} in the 'passive channel'. Thus one input is used for generating ramp voltage output waveforms; this input is referred to as 'virtual earth' and the other is utilized only for balancing and adjusting the ramp slope to zero, when the voltage applied to the integrating resistance is zero.

The use of a differential first stage with 2N2060's yields a considerable reduction of drift e_{D1} in the amplifier, originating predominantly from this stage during the scan period.

It has been shown by R. H. Okada[3] that transistors in differential stages of d.c. amplifiers should be matched for V_{be} and for current gain. These conditions are best satisfied by double planar epitaxial transistors, but it is also of importance that all resistors affecting their bias should have the same thermal coefficients to maintain steady base potentials and currents througout the operational temperature range and that the components determining the integration time-constant must have as low thermal coefficients as possible. Thus R_{29} and R_{30} should have the same thermal coefficients and R_1; RV_1 and C_8 should have as low thermal coefficients as possible. The bias stabilization for the differential stages of the amplifier is achieved through a form of a voltage stabilization of the linked third stage emitters, which is performed by VT_3. Although this transistor controls the emitter currents of the first stage, it does that more for the purpose of parallel amplification of collector voltages of VT_2 than for keeping the sum of their currents constant. The voltage of the linked emitters of VT_4 and VT_5 differs by their base-emitter voltages from the collector voltages of $VT_{2a,b}$. This emitter voltage, stepped down by resistors R_{11} and R_{12} is applied to the base of VT_3, which in turn compares it against the stabilized voltage of its emitter. Thus the loop is closed with VT_1 and VT_2 not contributing to the voltage gain at all.

For the single ended output stages, bias stabilization is provided by the overall voltage negative feedback through resistors R_{28} and R_{23} to the emitter of VT_{2b}. As the integrating capacitor C_8 is providing the negative a.c. feedback, capacitor C_7 is added to form a low-pass feedback through R_{28}, C_7 and R_{23} for reduction of the static drift.

For the best linearity the output compound emitter-follower stage, consisting of VT_{13}, VT_{14}, VT_{15} and VT_{16}, should operate in class-A under normal load conditions. The quiescent current through VT_{15} and VT_{16}, which depends on the voltage between the bases of VT_{15} and VT_{13}, should then be chosen relatively high within the transistor dissipation limits. As the emitter to base voltages are functions of temperature, for safety reasons it is better to include in the circuit compensating diodes MR_{10} and MR_{11}, which prevent this quiescent current from rising dangerously at high temperatures. R_{42} helps to maintain current through these diodes at negative output waveforms when VT_{11} and VT_{12} conduct small currents only.

In order to achieve satisfactory stability in this amplifier it is necessary to use transistors with high β cut-off frequency and to introduce phase correction networks operating at comparatively low corner frequencies. Since the integrator response has a (6dB/octave) negative slope beginning at $f = 1/(RC(1+G_1))$ the phase correction and stability do not present a serious problem.

The higher the β cut-off frequency of the transistors, the higher the corner frequencies should be chosen for the stabilizing networks, but for an integrator this choice is not critical and sufficient margin of stability can be obtained. Apart from C_5 in series with R_{22} and C_{11} with R_{40} (which are returned to earth) C_2 is added in series with R_7 (between the collector and base of VT_4) reducing the gain of the amplifier at high frequencies. The interstage network of R_{26}, C_{10} and L_1 in parallel with the base input impedance of VT_{11} utilize the flat series resonance effect and help to shape the amplifier response. In addition a 'lead' capacitance (not shown in the circuit) might be connected across R_{35} in order to reduce ringing at the beginning of the flyback period.

Diodes MR_6, MR_7 and the Zener diode MRZ_2 (3·3V) limit the amplitudes of the output voltage waveforms prior to their current amplification by VT_{10}. The action of other diodes MR_8 and MR_{12} is similar to that of MR_{10} and MR_{11}. Their purpose is to compensate the effect of variation of the base to emitter voltage with temperature by balancing it with that of a conducting diode in order to reduce the drift caused by single ended amplifiers. Thus VT_{10} is balanced by MR_8 and VT_{11} by MR_{12}. One more compensation should be mentioned here and it is that of the storage charge in the planar switch. This charge released directly into the integrating capacitor C_8 causes an undesirable positive going pedestal in the output voltage waveform and it must be compensated by the application through C_1 of a portion of gate voltage to the virtual earth. As the storage charge of a transistor increases with temperature so its compensating voltage is also varied correspondingly by the changing voltage across the conducting diode MR_4. Care should be taken not to allow the step of the gate, which passes through C_1 and C_8 to cause an excessive voltage spike in the integrator output voltage waveform due to S_1 being cut-off after a delay through the gate amplifier A_3. It may be necessary either to slow down the rise time of the gate, or to increase the load capacitance across the output of the integrator. By such means the output ramp voltage can remain satisfactorily stable within a range of ambient temperature from 20°C to 70°C.

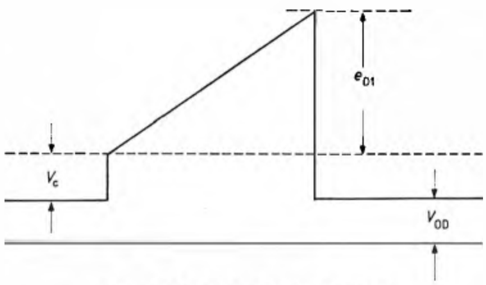


Fig. 3. Output voltage waveform

Simple Method of Setting-up the Integrator

Fig. 3 shows the integrator output voltage waveform for zero integrator voltage. Thus to set up the integrator it is necessary to short-circuit the integrator input terminals and then alter the system controls until the integrator output voltage is zero. This is achieved by adjusting RV_2 to remove V_c, adjusting RV_3 to remove eD_1, finally adjusting RV_4 to eliminate V_{OD}. The desired voltage to be integrated can now be introduced at the input terminals.

Conclusion

A ramp voltage having good linearity can be achieved by the use of an amplifier having high gain and relatively low bandwidth. The reset system, consisting of a limiting amplifier and two switches produces only a secondary error in the ramp linearity. This low error is attributable to the high off/on resistance ratio of the switches. This reset system also offers very good flyback characteristics and it can be used if very large scanning voltage amplitudes are required. Thus only the 'parabolic' error due to the limited gain of the inegrating amplifier is of significance and of the order of 2×10^{-4} at the gain $G_1 = 74$dB. The limited bandwidth of the integrating amplifier causes only a small delay in the start of the ramp voltage. This delay is approximately inversely proportional to the gain bandwidth product.

Acknowledgments

The authors wish to thank Messrs. Robinson and Lee for their support, and Dr. Scarr of S.T.C. for his useful suggestions. They are grateful to the directors of Cossor Electronics Ltd for permission to publish this article.

Appendix

(1) Reset System

During the scan period, the reset system as viewed from the input of the main amplifier (A_1) appears as a resistance R_{1F} in series with an e.m.f. of $R_{2N} E_L / R_{20}$, provided the reset amplifier (A_2) is operating in its limited mode.

Applying Kirchhoff's Law at the input to the main amplifier to give:

$$\frac{E - (V_o/G_1)}{R} = \frac{(R_{2N}/R_{20}) E_L + (V_o/G_1)}{R_{1F}} + (V_o/G_1 R_1) + (1 + (1/G_1))\, C (dV_o/dt);$$

solving this equation for V_o and making an approximation that at $t = 0$, $V_o = 0$ to give:

$$V_o = G_1 r \left(E/R - \frac{R_{2N} E_L}{R_{20} R_{1F}} \right) \left[1 - \exp\left(- \frac{t}{G_1 C_M r} \right) \right] \quad \ldots\ldots\ldots\ldots (6)$$

where $(1/r) = (1/R_1) + (1/R_{1F}) + (1/R)$ and

$C_M (1 + (1/G_1))C$

Expanding the exponential term in equation (6), using the equation for a perfect integrator as $V_{op} = (Et/CR)$, and putting $C_M \simeq C$ to give:

$$(V_o/V_{op}) = \left(1 - \frac{R_{2N} R E_L}{R_{20} R_{1F} E} \right) \left[1 - \frac{t}{2CrG_1} + \ldots \right] \quad \ldots\ldots\ldots\ldots (7)$$

If $G_1 \to \infty$

$$\frac{V_{op} - V_o}{V_{op}} \longrightarrow \frac{R_{2N} R E_L}{R_{20} R_{1F} E} \quad \ldots\ldots\ldots\ldots (8)$$

Equation (7) represents the error in the output due to the reset system.

(2) Reset System Limiting

For cases where at the end of the scan period the reset system is limiting, capacitor C will discharge at a nearly constant current through the reset circuit.

Assuming a virtual earth exists at the input of the main amplifier A_1, that $R_{1N} + R_{20} \ll R_{2F}$ that $\left| \dfrac{E_L}{R_{1N} + R_{20}} \right| \gg |E/R|$ and that R_{20} is much higher than the effective resistance between the integrator output and earth, then:

$$\frac{E_L}{R_{20} + R_{1N}} \simeq C\,(dV_o/dt) \quad \ldots\ldots\ldots\ldots (9)$$

Solving this equation and applying the boundary condition (for the case of a positive integrator input voltage) yields:

$$\tau = C\,(R_{20} + R_{1N})\,(E/E_L) \text{ provided } EG_{20} \gg E_L \;\ldots. \quad (10)$$

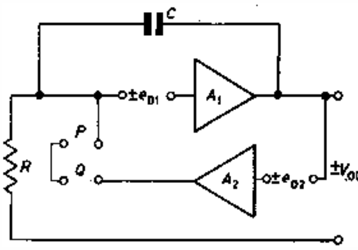
When the reset system ceases to limit the nearly linear fall off the output voltage changes to an exponential one and the equation for further discharge is:

$$(E/R) = S(CV_o) + (G_2 V_o / Z_T) \quad \ldots\ldots\ldots (11)$$

Where Z_T is the Thevenin impedance of the reset system viewed from the virtual earth and S is the Laplace transform. When the capacitor C is fully discharged $S(CV_o) = 0$. By replacing Z_T with $R_{1N} + R_{20}'$, and G_2 with G_{20} then from equation (11):

$$V_{OM} \simeq \frac{E(R_{20}' + R_{1N})}{G_{20}} \quad \ldots\ldots\ldots\ldots (12)$$

When the inut voltage to the integrator is sufficiently small not to cause the reset system to limit, the flyback response will then be determined directly by equation (11).



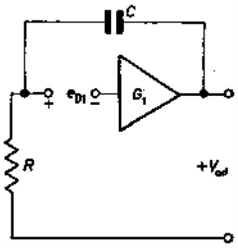


Fig. 4 (left). Equivalent circuit for output voltage at end of flyback period
Fig. 5 (right). Equivalent circuit during scan period

(3) INTEGRATOR OUTPUT VOLTAGE

(a) *Calculation of the integrator output voltage at the end of the flyback period due to drifts only in the A_1 and A_2 amplifiers.*

Assume C fully discharged, $R_{1N} \ll R_{20}' \ll R_{1F}$ and $R_{20}' \ll R$. The circuit may be presented as shown in Fig. 4.

When the reset loop is broken at points P and Q, the potential at P is: $\pm V_{OD}/G_{10} \pm e_{D1}$ and that at Q is: G_{20} $(\pm V_{OD} \pm e_{D2})$. When the loop is closed these potentials are the same and

$$V_{OD} = \mp e_{D2} \pm e_{D1}/G_{20} \text{ provided } G_{20}\,G_{10} \gg 1 \quad \ldots\ldots (13)$$

(b) *Calculation of the integrator output voltage during the scan period due to drifts only in the A_1 and A_2 amplifiers.*

Assume that amplifier A_1 takes no input current and that the reset loop is broken (between P and Q) (Fig. 5.)

Equating currents in R and C

$$\frac{e_{D1} - (V_{OD}/G_1)}{R} = CS\,[V_{od}\,(1 + (1/G_1)) - e_{D1}] \quad \ldots\ldots (14)$$

Solving this equation and applying boundary condition that at $t = 0$ yields:

$$V_{od} = e_{D1}\,G_1 + (V_{OD} - e_{D1}\,G_1)\exp\left(-\frac{t}{CR\,(1 + G_1)}\right) \quad \ldots\ldots (15)$$

expanding the exponential term and assuming $t \ll CRG_1$ and $G_1 \gg 1$ gives

$$V_{od} = e_{D1}\,t/CR - V_{OD} \quad \ldots\ldots (16)$$

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A Transistor Tester

By M. M. Winn*

A simple transistor tester is described. The transistor is tested at a collector current which is selected by the operator and the result is shown directly on a meter. The testing condition closely resembles the operation of a transistor in a switching circuit.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

SOME transistor testers use bridge methods to measure small signal parameters and therefore require manual balance of the bridge each time a transistor is tested. For rapid testing of transistors a direct reading feature is desirable. This is provided in a few transistor testers in which a direct current is injected into the base of the transistor and the collector current is shown on a meter calibrated in terms of the large signal current transfer ratio. Testers of this second type are simple and cheap. They are suitable for the sorting of switching transistors which should be classified according to the values of large signal parameters. However, the second type of tester performs the measurement at a controlled base current.

A test at a controlled collector current would be more desirable and this is performed in a circuit which is described below in which the collector current is chosen by the operator. In addition, the result is shown directly on a meter. The testing situation closely resembles the operation of a transistor in a switching circuit.

* *University of Sydney, Australia.*

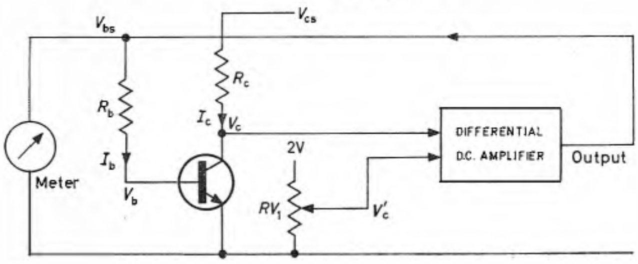
Method of Testing

The basis of the method is shown in Fig. 1. The transistor is placed in a feedback loop so that V_c is kept close to V_c' and by varying V_c' the transistor can be held in a controlled state of saturation.

Fig. 1. Basis of the method

$$I_c = \frac{V_{cs} - V_c}{R_c} \simeq V_{cs}/R_c \; : \; I_b = \frac{V_{bs} - V_b}{R_b} \simeq V_{bs}/R_b \; :$$

$$h_{fe} \simeq I_c/I_b \simeq \frac{V_{cs}R_b}{V_{bs}R_c}$$



The collector current:

$$I_c = \frac{V_{cs} - V_c}{R_c} \simeq V_{cs}/R_c \qquad \text{.......... (1)}$$

can be controlled by varying R_c.

$$I_b = \frac{V_{bs} - V_b}{R_b} \simeq V_{bs}/R_b \qquad \text{......... (2)}$$

and h_{FE} will be given approximately by:

$$h_{FE} \simeq I_c/I_b \simeq \frac{V_{cs}R_b}{V_{bs}R_c} \qquad \text{.......... (3)}$$

and for a given R_b/R_c set on switches h_{FE} can be read on a meter recording V_{bs}. (V_{cs} is kept constant.) As h_{FE} is proportional to $1/V_{bs}$ an inverse proportional meter scale is used. The inaccuracies incurred in equations (1) to (3) are fairly large especially if V_{bs} is small, but they are low enough for the purpose of rapid inter comparison of transistors. The ganging of the switches in the practical circuit facilitates the determination of the dependence of h_{FE} on I_c.

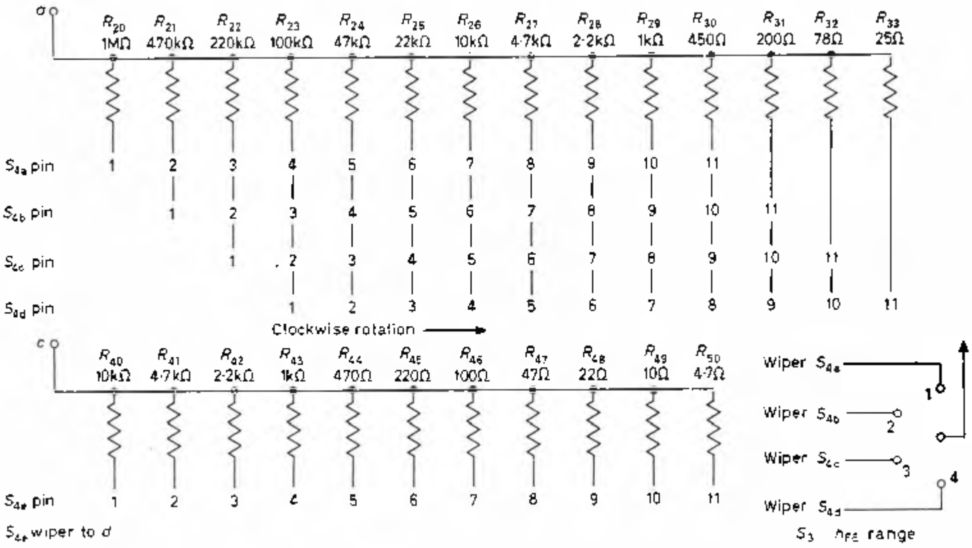


Fig. 2. h_{FE} meter arrangement

RANGE (1)	RANGE (2)	RANGE (3)	RANGE (4)	CORRESPONDS TO
10	20	50	100	f.s.d.
20	40	100	200	0·5 f.s.d.
50	100	250	500	0·2 f.s.d.

Practical Circuit

This is shown in Fig. 2. (Resistors are $\frac{1}{4}$W unless otherwise stated.) h_{FE} is measured in the following four ranges selected by S_3:

The value of I_c may be set by the 11 position switch S_4 to be any one of the following values (in milliamperes): 0·5, 1, 2, 5, 10, 20, 50, 100, 200, 500 and 1 000. The state of saturation is controlled by the potentiometer RV_1 which sets the voltage V_c' to any desired value between 0 and 2V.

The leakage current of the transistor may also be measured. This is done with the base open-circuited. The resistors R_8, on the 12 position switch S_5 provide a measure of protection for the meter and transistor during this measurement. Values of the meter shunts R_7, selected by S_5, give full scale deflexions (in milliamperes) of: 0·1, 0·2, 0·5, 1, 2, 5, 10, 20, 50, 100, 200 and 500.

Transistors may safely be plugged in and out of the test socket while the circuit is energized providing the I_c control sets a safe current and the type switch S_1 is correctly set.

The prototype instrument has been found quick and easy to use.

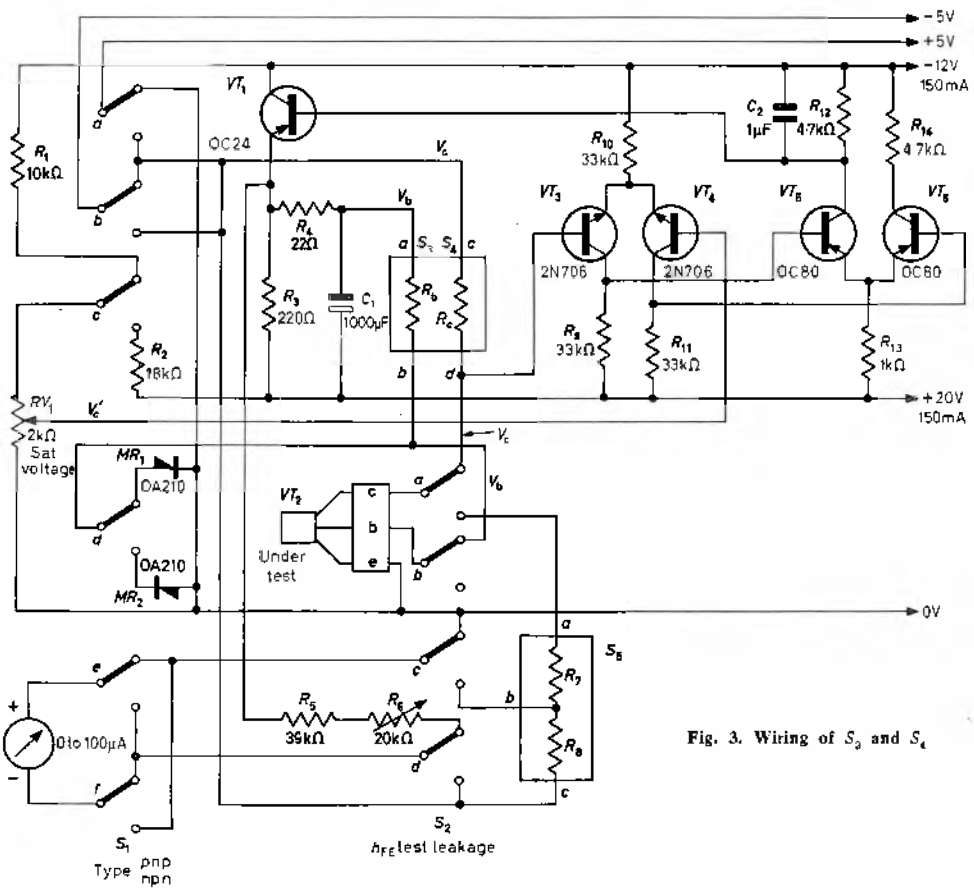


Fig. 3. Wiring of S_3 and S_4

Acknowledgments

This work has been supported by the Nuclear Research Foundation within the University of Sydney and by the Air Force Office of Scientific Research under Grant AF-AFOSR-305-63.

An Electron Current Amplifier

for use with an Electron Microprobe Scanning System

By J. P. Davey*, M.A., Ph.D.

The fundamental factors determining signal-to-noise ratio in an amplifier of small secondary electron currents are discussed, and it is shown that, in certain circumstances, there may be no significant advantage in the use of electron multiplication before amplification. A practical design is given for an amplifier of currents down to $10^{-10}A$ at bandwidths from d.c. up to 10kc/s, and the performance of this amplifier is described.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

THE general principles governing the design of amplifiers to attain the optimum signal-to-noise ratio with moderate values of source impedance are well established, but the problems arising when the source impedance is very large have, apparently, been studied only in special contexts (e.g. White and Harker[1], Gillespie[2]), and the conclusions that have been obtained are less widely known. In particular, in such systems as the electron scanning microscope[3] or the X-ray scanning microanalyser[4], in which a finely focused electron beam (electron microprobe) is scanned over the surface of a specimen, and the secondary electron current is used as a video signal to obtain a magnified image of the specimen surface, it has been usual to incorporate low-noise electron multiplication (using an open electron multiplier, or a scintillator followed by a photomultiplier) before amplification of the secondary electron current. In some circumstances the size of such an electron multiplying system can give rise to difficulties in the design of the specimen chamber, and it is therefore of interest to consider if an adequate performance can be obtained by direct amplification of the secondary electron current, which may then be collected by a simple electrode arrangement.

Sources of Noise in a Current Amplifier

The treatment here is similar to that given by Gillespie[2]. His analysis, however, deals with amplifiers intended for use with particle counters, the input signal consisting of a voltage step produced by the sudden arrival of a number of charged particles at the signal electrode; the results are therefore not in a form directly applicable to the present problem.

Gillespie found that the important sources of noise were thermal noise in the input resistor, grid-current shot noise, flicker noise, and anode-current shot noise. In the present context, another term must be added: the intrinsic shot noise, due to the inherent fluctuations in the current to be amplified, which arise from its composition as a sequence of discrete particles. It will here be assumed that their arrival at the collector is completely random, so that the intrinsic noise may be calculated by Schottky's theorem. This probably underestimates the intrinsic noise, since the current being amplified arises from secondary emission, and fluctuations both in the primary beam intensity and in the secondary emission coefficient contribute to those in the current collected. Space-charge smoothing, which might tend to reduce the noise level, is negligible in practice. The effect of an underestimate of the intrinsic noise is that the practical amplifier may be nearer to the ultimate signal-to-noise ratio than the present treatment suggests.

Equivalent Circuit of Input Stage

The equivalent circuit at the input of the amplifier is shown in Fig. 1, in which the noise sources are represented by current and voltage generators. I_1 represents the current to be amplified, i_1 the shot-noise fluctuations in this current (intrinsic noise), i_2 the thermal noise, I_g the grid current of the input valve, and i_3 the shot-noise fluctuations in this current. The internal noise sources in the valve, which produce fluctuations in its anode current in the absence of any voltage at the grid, are represented by two voltage generators, v_1 and v_2, representing flicker and shot

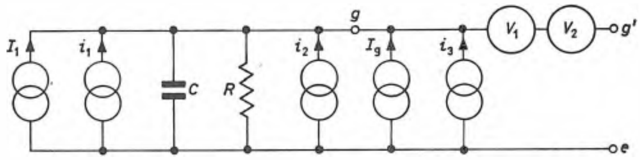


Fig. 1. Equivalent circuit of input stage

noise respectively, connected between g, the actual grid terminal, and g', the grid terminal of a fictitious noise-free valve. The capacitance C is the total capacitance seen at the input when the anode of the valve is earthed. It can be shown[2] that the apparent increase in input capacitance due to Miller effect does not change the signal-to-noise ratio.

Magnitudes of Noise Terms

The magnitudes of the noise contributions for an element of bandwidth df are given by:

$d\langle i_1^2\rangle = 2qI_1\,df$, (Schottky[5])

$d\langle i_2^2\rangle = 4kTR^{-1}\,df$, (Nyquist[6])

$d\langle i_3^2\rangle = 2qI_g\,df$,

$d\langle v_1^2\rangle = Af^{-1}\,df$, (Gillespie[2])

and $d\langle v_2^2\rangle = 4kTR_e\,df$, (van der Ziel[7])

where q is the electronic charge, k is Boltzmann's constant, T is the ambient temperature, and A is a constant which, for valves with oxide-coated cathodes, is approximately 10^{-13}. $\langle i_1^2\rangle$ etc. represent mean square values. R_e is given by the formulae:

$$R_e = \epsilon/g_m \quad \text{for a triode,}$$

$$R_e = \epsilon/g_m \cdot \frac{I_a}{I_a + I_s} + 20/g^2_m \cdot \frac{I_aI_s}{I_a + I_s} \quad \text{for a pentode,}$$

the coefficient ϵ varying from 2·5 to 4, according to the valve structure and operating conditions. It is assumed that the operating frequency is low enough for transit-time effects to be neglected, and that the valve is operated with its grid sufficiently negative to prevent its receiving electrons from the cathode.

The two voltage noise terms, v_1 and v_2, may be represented by equivalent currents i_4 and i_5, which, if flowing in the input circuit, would produce the same effect at the output.

* *Cavendish Laboratory, University of Cambridge.*

These are given by:

$$d\langle i_4^2\rangle = (R^{-2} + \omega^2 C^2)\, A f^{-1}\, df,$$

and

$$d\langle i_5^2\rangle = (R^{-2} + \omega^2 C^2)\, 4kT\, R_e\, df.$$

Because there is no correlation between any of these terms, the total equivalent noise current, referred to the input, is obtained by adding the mean square values of the individual contributions:

$$d\langle i^2\rangle = d\langle i_1^2\rangle + d\langle i_2^2\rangle + d\langle i_3^2\rangle + d\langle i_4^2\rangle + d\langle i_5^2\rangle.$$

NOISE RATIO

It is now possible to define a noise ratio:

$$N = \frac{\text{(Total noise power at output)}}{\text{(Intrinsic noise power at output)}}$$

analogous to that used to describe the performance of amplifiers which are limited by thermal noise[7]. In the present case, the fundamental limitation is the intrinsic noise term i_1; hence:

$$N = (d\langle i_1^2\rangle + d\langle i_2^2\rangle + d\langle i_3^2\rangle + d\langle i_4^2\rangle + d\langle i_5^2\rangle)/d\langle i_1^2\rangle$$

$$= 1 + (2kT/qI_1R) + (I_g/I_1) + \frac{(R^{-2}+\omega^2C^2)\,(Af^{-1}+4kTR_e)}{2qI_1}$$

It can be seen that:

(a) N is a function of current. If the noise performance is satisfactory at the lowest value of input current I_1, it will be satisfactory at all higher values.

(b) The thermal noise contribution is inversely proportionally to R. It can be made negligible if $R \gg 2kT/qI_1$.

(c) N is dependent on frequency. The flicker-noise contribution contains the term $A/2R^2qI_1f$, which becomes infinite for $f = 0$. A similar term occurs in the expression for noise in any direct-coupled amplifier, and may be regarded as a component of the long-term drift. Drift may be excluded by imposing an arbitrary low-frequency cut-off (e.g. at 0·1c/s), when the divergent term becomes negligible. In the remaining analysis this term, and the practically unimportant R^{-2} shot noise term will both be neglected. The other flicker and shot noise terms increase with frequency, so that the effective noise ratio of a wide-band amplifier depends on its overall frequency response. In particular, since the shot-noise term varies as f^2, the high-frequency cut-off must be steeper than 6dB/octave if the total noise is to be finite.

Overall Frequency Response of System Used

If the output of the amplifier is used, as in the present work, to modulate the beam intensity of a cathode-ray tube, then the effective bandwidth is often determined by the scanning speed and width of the focused spot. It can be shown that, if the current density distribution in the spot is of the Gaussian form $\exp(-r^2/2r_0^2)$, and the spot moves with velocity v, then the response to a sinusoidal modulation varies as $\exp(-\omega^2/2\omega_0^2)$, where $\omega_0 = v/r_0$. Measurements made on an actual cathode-ray tube indicate that this law is followed with reasonable accuracy.

TOTAL NOISE

If it is assumed that the overall response is of this Gaussian form, then the total fluctuation referred to the input is given by:

$$\langle i^2\rangle_t = \int_0^\infty \exp(-\omega^2/\omega_0^2)\,(d\langle i^2\rangle/df)\,df$$

$$= \omega_0(qI_1 + 2kTR^{-1} + qI_g)/2\sqrt{\pi} + \tfrac{1}{2}C^2(A\omega_0^3 + kTR_e\omega_0^3/\sqrt{\pi}).$$

An average noise ratio N' can be defined for the wide-band amplifier in a similar manner to N for the narrow-band amplifier:

$$N' = 1 + (2kT/qI_1R) + (I_g/I_1) + (C^2/qI_1)(A\omega_0\sqrt{\pi} + kTR_e\omega_0^2)$$

The signal-to-noise ratio S may also be calculated:

$$S = I_1/\sqrt{\langle i^2\rangle} = \frac{(2\sqrt{\pi}I_1)^{\frac{1}{2}}}{(Nq\omega_0)^{\frac{1}{2}}}$$

It can be seen that the required bandwidth and signal-to-noise ratio set a lower limit for I_1, even if a perfectly noiseless amplifier can be used; appreciable benefit from the use of electron multiplication is likely only if a low value of N' cannot be achieved with an amplifier of appropriate bandwidth. An arbitrary criterion for the range of usefulness of direct amplification is that $N' \leqslant 4$; the signal-to-noise (voltage) ratio will then not be worse than half that of an ideal system. This can be achieved if:

$$I_g + (2kT/qR) \leqslant I_1 \text{ and } C^2(A\omega_0\sqrt{\pi} + kTR_e\omega_0^2)/q \leqslant 2I_1.$$

A better noise ratio is obtained when the grid current and anode current shot noise terms are equal, but since the grid current, which is usually somewhat unstable, flows in the input circuit, the drift can be reduced by working at a lower value of grid current, with reduced bandwidth.

Practical Limits of Direct Amplification

The valves which are here considered for use in the first stage are the Mullard types E88CC (one section), ME1400, and ME1403. The characteristics are taken from the manufacturer's data, supplemented by the author's measurements. R_e for the ME1400 and ME1403 is calculated from the formulae of van der Ziel, using $\epsilon = 4$. The ME1400, which is a pentode, is assumed to be triode connected, as a lower value of R_e is thus obtainable. The use of triode connexion for the ME1403 would result in a first-stage gain of only 6dB, and the second-stage noise would then become important. The increased R_e for this valve in the pentode connexion is, however, less significant, because of the relatively greater importance of flicker noise at low frequencies. The input capacitance C is assumed to be 30pF, a value which can be achieved with reasonable care in the construction of a practical amplifier.

For a given value of the input current I_1, it is possible to select appropriate conditions for the input valve. When this is done, a value of R may be found for which thermal noise is nearly negligible. The characteristic frequency $f_0\,(=\omega_0/2\pi)$ for which $N' = 4$ can then be determined, and the expected value of the signal-to-noise ratio S can

TABLE 1
Operating Conditions of Input Valve

VALVE TYPE	V_a (V)	I_a (A)	V_g (V)	I_s (A)	g_m (A/V)	I_g (A)	R_e (Ω)	I_1 (A)	R (Ω)	f_0 (c/s)	S
ME1403	10	5μ	6·5	2·2μ	10·5μ	3×10^{-15}	550k	10^{-14}	10^{13}	2·9	56
ME1403	10	100μ	10	60μ	100μ	4×10^{-14}	100k	10^{-13}	10^{12}	25	59
ME1403	10	100μ	10	60μ	100μ	4×10^{-14}	100k	10^{-12}	10^{11}	130	81
ME1400	45	100μ			300μ	5×10^{-12}	13k	10^{-11}	10^{10}	1·1k	90
ME1400	50	500μ			950μ	$1{\cdot}2\times10^{-11}$	4·2k	10^{-10}	10^{9}	7·2k	110
E88CC	90	3m			5m	$4{\cdot}5\times10^{-10}$	1·2k	10^{-9}	10^{8}	41k	150
E88CC	90	10m			10·5m	$1{\cdot}6\times10^{-9}$	400	10^{-8}	10^{7}	250k	190
E88CC	90	10m			10·5m	$1{\cdot}6\times10^{-9}$	400	10^{-7}	10^{6}	820k	330

be estimated. Typical values thus obtained are given in Table 1.

It can be seen that the use of direct amplification in scanning microscopes (which require bandwidths of a few kc/s) is restricted to those instruments in which the minimum secondary electron current is rather greater than 10^{-11}A. In the author's case, when the minimum probe current was of the order of 10^{-10}A, it was decided to use direct amplification. The circuit used is described below.

A Practical Amplifier Design

Of the possible operating conditions for the input valves listed in Table 1, the ME1400, operated at 500μA anode

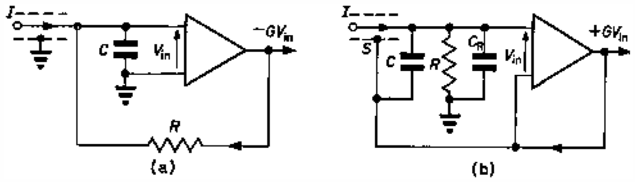


Fig. 2. Application of negative feedback

current, was chosen as the most suitable. The value of grid resistor required, at least $10^9\Omega$, presents a problem, since the impedance of the shunt capacitance falls below this at frequencies greater than 5c/s. To obtain an overall flat response it is therefore necessary to use some form of frequency correction, as in television camera amplifiers[1,8]. Because of the very wide variation of input impedance over the pass-band, the use of passive correction at a late stage in the amplifier may result in the earlier stages being overloaded by the low-frequency components of the signal. It is better to use negative feedback to provide the frequency correction; two ways of doing this are illustrated in Fig. 2.

Effect of Negative Feedback

In the circuit of Fig. 2(a), which may be termed the virtual earth connexion, shunt feedback is applied via the grid resistor R. The effective input resistance of the amplifier (of gain G) is reduced to R/G, the capacitance being unchanged. In Fig. 2(b), the signal lead is surrounded by a shield electrode S, which is driven from the amplifier output in such a way that its potential closely follows that of the input connexion I. By this means, the effective capacitance between the input lead and the shield is reduced to C/G. However, there will be some irreducible capacitance between the signal electrode and earth, represented by C_R, and the minimum effective input time-constant obtainable is thus RC_R. This limitation does not apply to the virtual earth connexion, which has the additional advantage that the screening electrode may be connected directly to earth.

The virtual earth connexion has been used in the present amplifier, the circuit of which is shown in Fig. 3. The feedback network actually used, which is similar to one used by Allenden[9] in an electrometer amplifier, is more complex than that of Fig. 2(a). The performance of the simpler circuit would be affected at high frequencies by the self-capacitance of R_1 and by other stray capacitance. In the present design, negative feedback takes place through C_1 at high frequencies, and the overall frequency response will be flat if $C_1R_1 = C_2R_2$. This adjustment is made by observation of the response to a square pulse input. When this is done, the feedback network is very nearly equivalent to that of Fig. 4, as may be shown by the star-delta transformation. It can be seen that there are insignificant additions to the shunt impedances at the input and output of the amplifier, while the effective capacitance shunting R_1 is reduced to a very low value. The trimmer C_3 enables this capacitance to be varied to compensate for high-frequency phase-shift within the amplifier, which might otherwise tend to produce positive feedback at frequencies outside the pass-band.

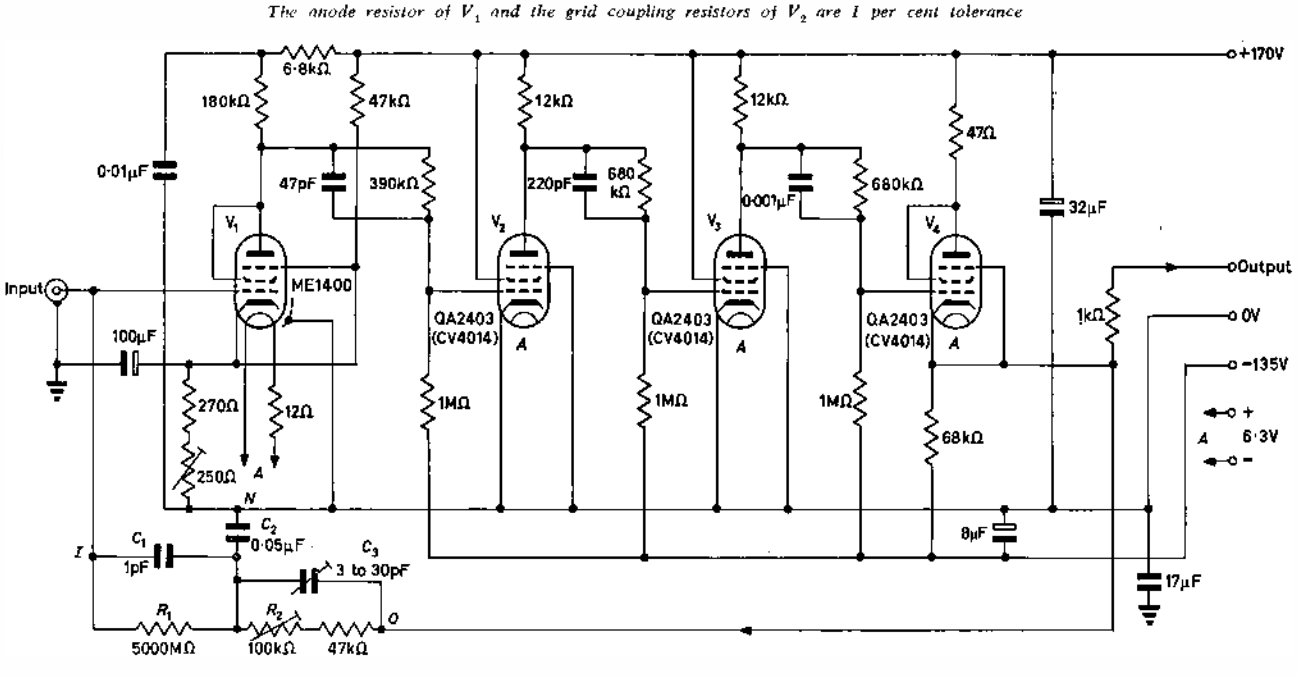
Performance

An amplifier with the circuit of Fig. 3 has been constructed, and has been used in the electron signal channel of an electron microprobe scanning system*. The amplifier

* *The system has been designed for the study of cathodoluminescence; a description of it is in course of preparation.*

Fig. 3. Circuit of complete amplifier

The anode resistor of V_1 and the grid coupling resistors of V_2 are 1 per cent tolerance



has a bandwidth of 25kc/s, and is linear for currents of up to 5×10^{-9}A, at which the output is 25V. It is used to modulate the beam of a cathode-ray tube, type 5BKP7 (Electronic Tubes Ltd), for which the characteristic frequency f_0 is approximately 4·5kc/s, under the usual operating conditions (in which a line 9cm long is scanned in 1/50sec). Under these conditions, the theoretical noise ratio is 2·2, and the signal-to-noise ratio 190 ($I_1 = 10^{-10}$A). The amplifier has adequate bandwidth to be usable at a line frequency of 100/sec ($f_0 = 9$kc/s), though the noise ratio then worsens to 4·7.

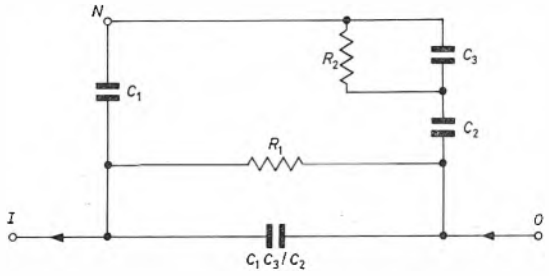


Fig. 4. Equivalent circuit of feedback network

Measurement of the signal-to-noise ratio actually obtained has proved difficult. The noise is not apparent on the image of a specimen (of fairly high intrinsic contrast), recorded in 10sec, the minimum time for which the structure of the raster can be made to disappear. An approximate visual estimate, made from observation of the waveform as seen on an oscilloscope (with the bandwidth restricted to 4·5kc/s), gave a signal-to-noise ratio of 80. Much of the difference from the theoretical value of 190 can be attributed to the fact that the simple filter used to restrict the bandwidth had a more gradual cut-off (12dB/octave) than the Gaussian frequency transfer characteristic assumed for the cathode-ray tube.

A practical problem in the use of the amplifier has been stray a.c. pick-up. This has been minimized by the use of a stabilized d.c. supply for the valve heaters, and by ensuring that there are no large transformers or similar devices in the neighbourhood of the amplifier. Some ripple was introduced by the variable (0 to 20V) bias supply, connected between the amplifier 0V line and earth, which was used to attract the secondary electrons to the collector electrode. If it were not required to vary this bias voltage, it would be preferable to obtain from it a small battery connected in series with the input lead.

Conclusion

The practicability of direct amplification of the secondary electron current has been demonstrated for a particular electron microprobe scanning system. No unusual problems have occurred in the construction, adjustment, or operation of the amplifier. The discrepancy between the theoretical and practical estimates of the signal-to-noise ratio may not be significant, in view of the difficulties of measurement, and the fact that the theoretical values for the valve noise contributions are averages, from which an individual valve may show considerable differences.

Acknowledgments

The author wishes to express his gratitude to his supervisor, Dr. V. E. Cosslett, for the help and encouragement given him throughout the course of his work, and to Dr. K. C. A. Smith and Dr. P. Duncumb, for helpful criticism of this article. He acknowledges financial support from the Department of Scientific and Industrial Research, and from N.V. Philips Gloeilampenfabrieken.

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A Portable Heart Beat Monitor

Although appearances may suggest that a man is dead he is no doubt still alive as long as his heart is beating and a doctor will never give up a victim of an accident if there are any indications at all of the heart still beating.

While today hospitals are well equipped with the most advanced instruments for clearly detecting even the slightest heart actions, the doctor who is giving medical assistance at the place of the accident has no other possibility than feeling the pulse or auscultating the chest in order to ascertain the patient's condition. This method, which is inadequate even when applied by experienced doctors, is the more insufficient when being used by the ambulance men, firemen or the patrol-car policemen, who frequently have to give first aid.

Teldix, of Heidelberg, has now developed a transistor unit the size of a pocket radio, which anybody can carry about with him and which permits the unambiguous detection of even the weakest heart beats. Short 3kc/s audio pulses sound in the loudspeaker in the rhythm of the heart beats. The volume may be regulated by a two-step switch. The loudspeaker volume can be adjusted to 'loud' or 'soft' depending on circumstances.

The technical principle of this unit is based on tapping off the small voltage differences caused by the heart beats, then amplifying them and finally converting them into an audible signal.

An impedance convertor stage forms the unit input. This stage is essential because the human body may be considered a high-impedance generator. Its internal resistance is determined mainly by the skin resistance, moreover, the transfer resistance between the skin and electrode must be borne in mind. However, since transistor inputs feature a low impedance as a rule, this convertor stage must be provided with an input resistance of approximately 200kΩ. This stage is followed by an electronic filter, in which interfering 50c/s voltages are filtered out. To this filter a four-stage amplifier is connected, which contains a low-pass filter having a out-off frequency of about 25c/s. The pulse voltage supplied by this amplifier actuates in the next stage a 3kc/s multivibrator, one of its arms being connected to a loudspeaker. In this manner each pulse caused by the heart beat is converted into a loud audible signal. The length of each signal is dependent on the pulse width and on the threshold at which the multivibrator is actuated.

The impedance convertor stage is provided with the low-noise transistor BSY72, and all other stages are fitted with the transistor 2N2926. Both types are passivated npn silicon planar transistors, which ensure long life and good temperature stability.

Power is supplied by two 3V batteries. One battery feeds the amplifier, the other the multivibrator. This arrangement was selected to prevent feedback between the various stages which could arise due to the high amplification factor. The unit measures only 135 × 95 × 33mm (without case) and weighs only 0·5kg with carrying case and associated electrodes.

A Microelectronic Frequency Divider with a Variable Division Ratio

By B. Preston*

The operation of a frequency divider system in which it is possible to set the division ratio to any desired whole number is described[1]. A more detailed circuit description is given with emphasis on the techniques required to produce the circuit in microelectronic form. Results and waveforms obtained from such a circuit are given.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

THIS article describes the development of a frequency divider system the division ratio of which can be varied as desired and one which may be translated readily into microelectronic form. The initial requirement for such a system was for inclusion in frequency synthesizers using the phase locked loop principle. It is envisaged, however, that such a system could be used in a variety of other applications such as highly accurate and stable timing sources, batch counters, standard pulse generators, including television synchronizing pulse generators, and in numerous frequency counting operations.

Frequency division can be achieved in a number of different ways. Microelectronic techniques cannot be readily applied to all of the methods available. Other systems fall short in the basic requirements for a frequency divider system with a variable division ratio. For example, a well tried and tested method of frequency division is the sinusoidal regenerative loop. This device requires up to three tuned circuits for each stage of division making it impossible to gain any advantage from microminiaturization.

A circuit that would lend itself to translation into microelectronic form and one which can also be designed into a frequency divider system with a variable division ratio is the locked astable multivibrator. The great disadvantage with this device is that it is not a fail safe system. There would always be an output from the device even if the input or any stage, apart from the final one, failed.

Perhaps one way to overcome the objections of a free-running multivibrator is to use a system of bistable devices. This would certainly provide a fail-safe device and also one admirably suited to microelectronics. However, the complexity of circuit and switching needed to produce an almost infinitely variable division ratio makes this system less desirable.

The system finally adopted uses as its basic building block a staircase frequency divider which utilizes the pulse-counting technique of frequency division. The majority of the failings listed in the above systems are eliminated, and with careful design of building blocks an economical unit can be produced.

Overall System

A pulse-counting frequency divider is a digital counting circuit to which input pulses are applied at a frequency f and which gives a pulse out for every n pulses in. The output frequency is thus f/n. Binary and decimal dividers have division ratios of two and ten respectively. Certain frequency synthesizers require dividers which can be set by decade dials to divide an input frequency in the megacycle range by any number between 1 and 999 or between 1000 and 1999, as selected by the dials[2].

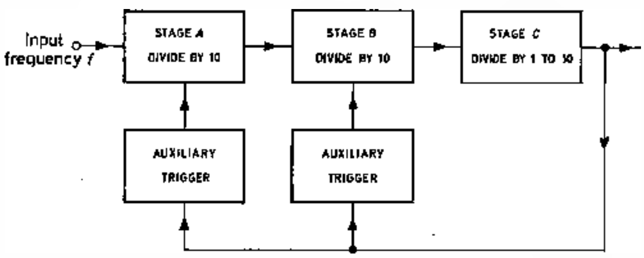


Fig. 1. Arrangement of frequency divider

Bearing in mind the objections mentioned in preceding paragraphs it will be realized that the production of such a device in microelectronic form requires an entirely new approach. The block diagram of such a frequency divider, where the division ratio can be set to any number between 1 and 999, is shown in Fig. 1.

It consists basically of three frequency divider stages A, B and C. Stages A and B have an inherent fixed division ratio of ten, but are capable of having this division ratio increased or decreased during any one particular count of either stage. This is achieved by bringing the auxiliary trigger circuit into operation, the length of time it operates and hence the number of pulses added to or subtracted from the standard count of ten being controlled by the dial setting. Stage C, on the other hand, is a frequency divider stage of similar construction but which is capable of having its division ratio set to any number between one and ten, where the division ratio set operates on every count of the stage, not being modified by any auxiliary trigger circuit.

To illustrate the operation of the overall circuit, consider the sequence of events that occur when a division ratio of 458 is required. The decade dial controlling the division ratio of stage C is set to 4, while the dials controlling the duration of operation of the auxiliary trigger

* *Microelectronics Division, The Marconi Co. Ltd.*

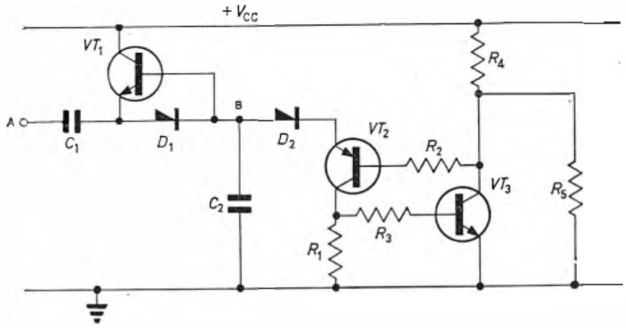


Fig. 2. Basic frequency divider

circuits associated with the stages A and B are set to 8 and 5 respectively. The durations refer to the period of the input pulses to the respective stages.

The auxiliary trigger circuits are initiated by a pulse appearing at the final output, stage C. The trigger circuits discharge the 'store' in their associated divider stages, without giving an output from that particular stage, after the period determined by the dial setting. Thus the count of stages A and B is modified from their usual count of 10 once every time there is a pulse from the final output.

To obtain one pulse at the final output requires four pulses from the output of stage B. In terms of output pulses from stage A these four pulses are made up of $(15 + 10 + 10 + 10) = 45$ pulses out of stage A. These 45 pulses are made up of one count of 18 input pulses to this stage and 44 counts each representing a period of 10 input pulses, thus giving the desired overall count of 458 input pulses to produce one output pulse.

It must be mentioned here that the very first count of the system after switching on the supply will be incorrect as there will have been no final output pulse to operate the auxiliary trigger circuits. In this case the initial count of the system would be one of 400.

Basic Frequency Divider

The circuit diagram of the basic frequency divider is shown in Fig. 2. The positive going leading edge of a pulse applied to input point A will cause diode D_1 to be forward biased and transistor VT_1 cut off. The peak amplitude of this positive edge will charge capacitors C_1 and C_2 to voltage levels dependent upon their relative values. On the negative going edge of the same pulse, diode D_1 will become reverse biased and thus, with transistor VT_2 cut off due to the more positive potential on its base set by resistors R_4 and R_5 will isolate capacitor C_2. Capacitor C_1 on the other hand, will fully discharge when transistor VT_1 conducts on the negative excursion of the pulse. Thus at the conclusion of the first input pulse, capacitor C_2 has accumulated a charge while capacitor C_1 is left fully discharged[3,4]. This process is repeated with each successive input pulse until the potential on C_2, and hence on the emitter of transistor VT_2, is more positive than the trigger potential on the base of VT_2, set by resistors R_4 and R_5. At this trigger point, transistor VT_2 will conduct, in turn bringing VT_3 into conduction and thence positive feedback causing a rapid discharge of capacitor C_2. Thus a staircase waveform is obtained at point B (Fig. 2). The base of VT_2 is now sitting at a potential equal to the bottoming potential of VT_3 plus the voltage drop across resistor R_2. When C_2 has discharged sufficiently to bring the potential on the emitter of transistor VT_2 to about its own base potential this transistor will cut off, which in turn will switch off transistor VT_3. The circuit is now back to its original state where the staircase waveform can again be formed at point B[5].

It will be noted that the circuit described gives a negative output pulse. This is ideal for direct coupling into a complementary version of the circuit (Fig. 3). This in turn produces a positive output pulse enabling further stages to be cascaded.

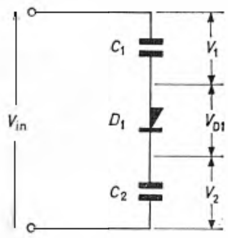


Fig. 4. Equivalent circuit of the input network of a divider stage

Referring to the equivalent circuit diagram of the input network of a divider stage (Fig. 4.)

$$V_{in} = V_1 + V_{D1} + V_2$$

$$\text{Let } C_2 = NC_1 = C$$

Then as the capacitors will obtain the same charge with each step

$$CV_1 = NCV_3 \text{ hence } V_1 = NV_2$$

$$\therefore V_{in} = V_{D1} + V_2(N + 1)$$

$$\text{and } V_2 = \text{Step voltage} = \frac{V_{in} - V_{D1}}{(N + 1)}$$

The trigger circuit (transistors VT_2 and VT_3) will turn off when capacitor C_2 has discharged to a voltage level equal to $V_{BE3} + V_{CE(sat)2} + V_{D2}$, where V_{BE} is the base emitter voltage of the transistor when conducting, $V_{CE(sat)}$ is the saturation voltage and V_D is the voltage across the diode when forward conducting.

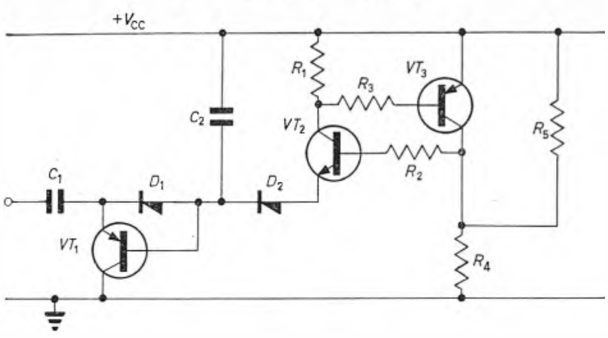
(The suffixes denote the transistors referred to.)

This potential therefore becomes the effective starting potential of the staircase waveform.

$$\text{The staircase trigger potential} = \frac{V_{CC}R_5}{R_4 + R_5} + V_{BE2} + V_{D3}.$$

(V_{CC} is the supply voltage).

Fig. 3. Complementary circuit



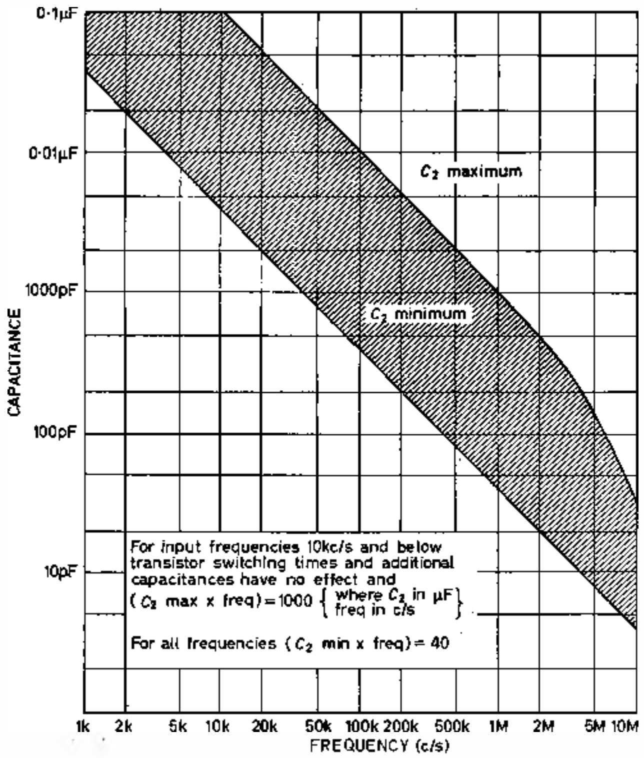


Fig. 5. Limit values for C_2

Assuming that the base-emitter voltage of a transistor when conducting is the same for VT_2 and VT_3, then the voltage available for the staircase waveform =

$$\frac{V_{CC}R_5}{R_4 + R_5} - V_{CE(sat)2}$$

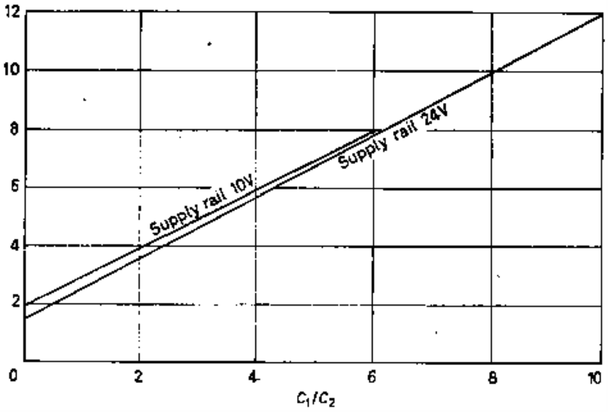
If the required division ratio is to be, say X, for maximum stability the trigger circuit is required to operate after $(X - \frac{1}{2})$ steps.

$\therefore$ Step voltage required =

$$\frac{V_{in} - V_{D1}}{(N+1)} = \frac{\dfrac{V_{CC}R_5}{R_4 + R_5} - V_{CE(sat)2}}{(X - \frac{1}{2})}$$

$$\therefore \text{Capacitor ratio } N = \frac{(X - \frac{1}{2})\ (V_{in} - V_{D1})}{\dfrac{V_{CC}R_5}{R_4 + R_5} - V_{CE(sat)2}} - 1.$$

Fig. 6. Capacitor ratio plotted against division ratio



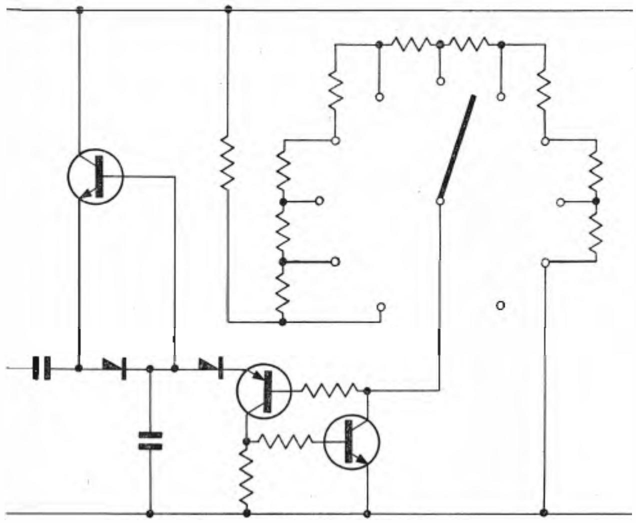


Fig. 7. Potentiometer method of altering the count of a divider stage (stage C)

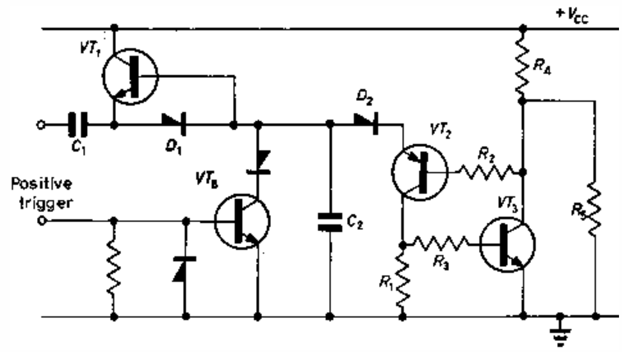


Fig. 8. Method of increasing the count in stages A and B

If the divider stage is being driven from a similar circuit the driving voltage $V_{in} = \dfrac{V_{CC}R_5}{R_4 + R_5} - V_{CE(sat)3}$

The above calculations determine the ratio of capacitors C_1 and C_2. The absolute values are dependent mainly upon the value of the resistor R_1 and the frequency of the input signal.

In the trigger circuit resistors R_1 and R_4 have minimum limits set by the peak current that can be handled by the transistors, while resistors R_2 and R_3 are inserted to limit the peak base currents in transistors VT_2 and VT_3 respectively.

Once the value of R_1 has been decided upon, the maximum value of C_2 is determined by the rate at which it is required to be discharged through VT_2 and R_1. To avoid an incorrect division ratio C_2 must be fully discharged before the next input pulse arrives. At the higher frequencies of operation the switching time of the transistors becomes a significant part of the inter-pulse period. Because of this it is necessary to have the time-constant C_2R_1 as small as one quarter of the period of the input pulses.

The minimum value of C_2 is governed by the amount of 'droop' or 'tilt' that can be tolerated during the waiting period between each voltage step. The droop is caused by the leakage current of transistor VT_2 partially discharging C_2 while tilt in the step is caused by the leakage currents of VT_1 and D_1 putting additional charge into C_2. It is fortunate that these two effects tend to cancel one another but it could be possible for one to become excessive especially at elevated temperatures.

When divider stages are cascaded the peak-to-peak drive available to each stage is fixed and constant. Under these conditions the range of possible values for C_2 can be obtained from Fig. 5 and the ratio of C_1 to C_2 from Fig. 6.

Divider Stage with a Variable Count (Stage C, Fig. 1)

A very convenient way to alter the count of a particular divider stage, without altering the ratio of the capacitors C_1 and C_2, is to vary the aiming potential of the trigger circuit, set by the potentiometer network (Fig. 7). By suitably adjusting the aiming potential it is possible to make the basic divider stage operate with division ratios of from say one to ten. The division ratio set, however, is fixed for every count of that stage, until reset to another number. Varying the aiming potential in this manner does, however, mean that the peak-to-peak amplitude of the output pulse will decrease as the division ratio decreases.

As stated previously this method of varying the count of a frequency divider is suitable for use in stage C of the overall system. For stages A and B it is necessary to have a circuit that will operate on the capacitor store, C_2, for one particular count, after which the stage must return to its inherent division ratio of ten. Also to ensure

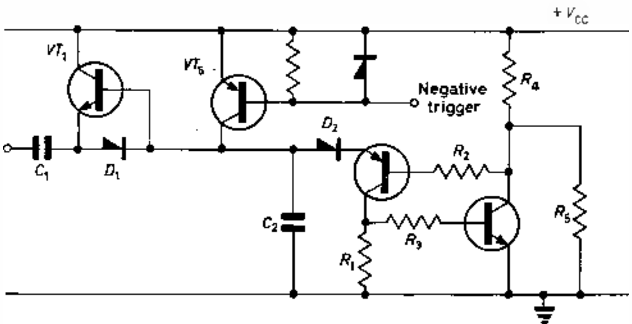


Fig. 9. Method of decreasing the count in stages A and B

that this divide by ten state is not changed in any way the capacitor store must be discharged to a potential similar to the normal starting potential of the staircase, that is $= V_{BE3} + V_{CE(sat)2} + V_{D2}$. A circuit that achieves this is shown in Fig. 8. It consists of a transistor VT_5 which is normally in the cut-off condition but which can be brought into conduction by applying a suitable positive going pulse. When this transistor conducts it rapidly discharges C_2, to a level set by the saturation voltage of the transistor itself plus the forward voltage of diodes in its collector, without operating the main trigger circuit. The divider stage then has to perform its normal count of ten before an output pulse is obtained, thus effectively increasing the division ratio for one count only.

If now a complementary version of the circuit, described above is also added to the standard frequency divider stage (Fig. 9) it gives a means of operating the main trigger circuits, so discharging C_2, at any point desired. This is achieved by letting transistor VT_4 conduct heavily for the duration of a negative pulse applied to its base, thus taking the potential on the emitter of transistor VT_2 more positive than its trigger point. Thus the division ratio can be reduced from an inherent divide by ten for any one particular count, the factor by which the count is

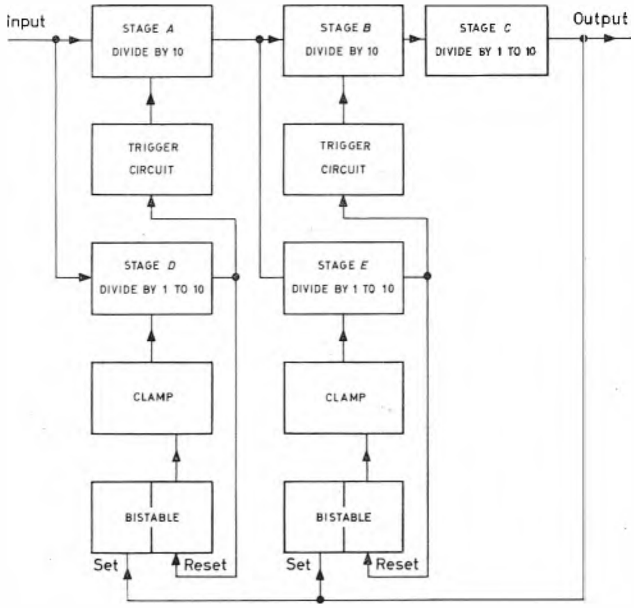


Fig. 10. Expanded block diagram

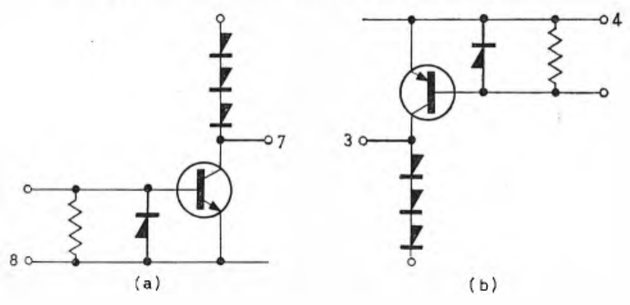
reduced being controlled by the time at which VT_4 is brought into conduction. This circuit also provides a very effective method of converting a divider stage with a normally fixed division ratio to one with similar characteristics to stage C. It has one advantage over the type used for stage C in that the peak-to-peak amplitude of the output pulse is constant irrespective of the stage division ratio, this being obtained of course at the expense of considerably more complex circuits.

The component parts of the auxiliary trigger circuit are expanded further in block diagram form in Fig. 10. The blocks marked 'Trigger Circuit' perform the function described in the preceding paragraph.

Divider stages D and E are similar in construction to stage C, the division ratio being variable from one to ten by varying the potential set at the collector of transistor VT_8 (Fig. 2). These two divider stages are, however, normally held inoperative by the action of the clamp circuit. This function is achieved by a transistor, normally held in the conducting state, placed across the store capacitor C_2. This makes the stage inoperative until the clamp is released by a change of state of the bistable. This method of control is adopted in order to retain uniformity of circuits and thus avoid having to gate the signal applied to the input of the stage.

The bistables controlling the state of the clamp circuits are set by a pulse from the final output. This change of state of the bistables releases the clamp circuits on stages

Fig. 11. Universal amplifier, trigger and clamp module



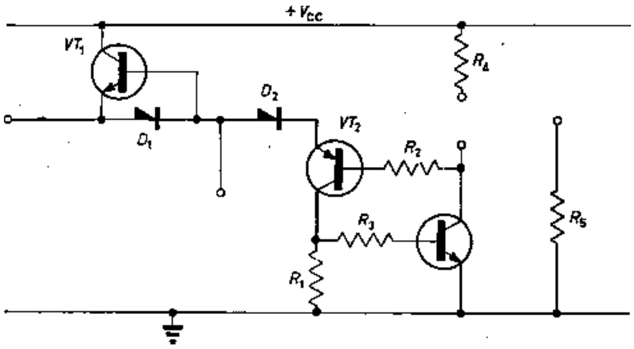


Fig. 12. A standard divider module

D and E and allows them to perform their preset count. The pulse obtained from the outputs of stages D and E then performs two functions. Firstly it operates the trigger circuit (Fig. 8) on the associated divider (stage A or B) causing the storage capacitor C_2 to discharge rapidly. Secondly it is fed back to reset its associated bistable, thus rendering divider stages D and E again inoperative until the next output pulse appears.

In the manner described above it is thus possible to modify the count of divider stages A and B for one count every time there is a pulse from the final output. All succeeding operations of these divider stages will have the inherent count of ten.

To translate the block diagram into a conventional circuit requires additional limiting amplifiers and several invertor stages. These amplifier stages would require to have their outputs referenced to either earth or the positive supply rail. A convenient way to produce such an amplifier in microelectronic form is to use two complementary class-C stages in one can. To obtain the desired fast rise times at the collector terminals necessitates the use of load resistors of 1kΩ or less. With a 24V supply rail this means a peak dissipation in excess of 570mW, which is far too great for a diffused resistor, making it necessary to use external components. As the trigger and clamp circuits also require both npn and pnp transistors, the module shown in Fig. 11 is used as a universal device for trigger and clamping circuits and, with external resistors between pins 7 and 4 or between pins 3 and 8, as a limiting or inverting amplifier module.

Fig. 13. Adding 5 pulses into the 'tens' column of an overall count of 655

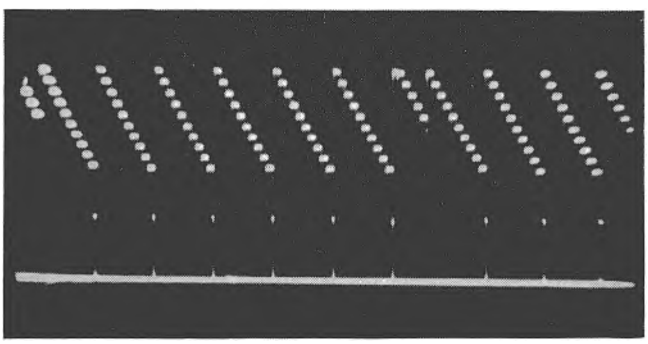
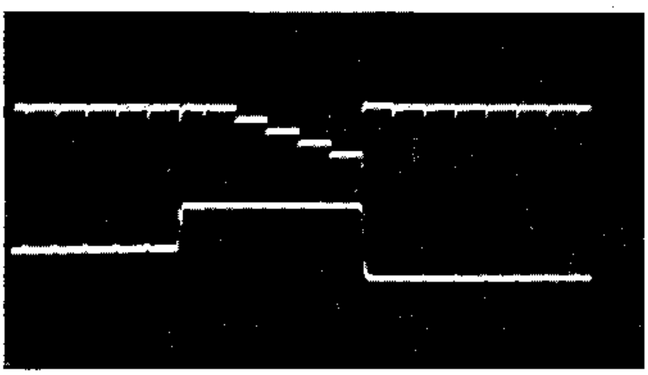


Fig. 14. Bistable output controlling pulse count of divider stage



This is part of the new approach the circuit design engineer has to adopt when producing circuits in microelectronic form. It is most important that as many common building blocks as possible are used even if it means sacrificing a small amount of reliability by using standard components external to the can[6].

As far as microelectronic building blocks are concerned a certain amount of standardization is achieved in the divider modules. The basic variables, in all the types of divider stages used, are the ratio of the capacitors C_1 to C_2 and the trigger potential set at the collector of VT_3. To produce a standard module the capacitors C_1 and C_2 are not included in the encapsulation and also, the collector of VT_3 and resistors R_4 and R_5 are taken to separate pins (Fig. 12). In this way the circuit designer has control over the frequency of operation and the division ratio set by C_1 and C_2, and also the facility of being able to use the standard module as a frequency divider with either a fixed or variable division ratio.

Again, in the interests of standardization, the bistable units are made identical. This entails standardizing on the polarity of the set and reset pulses. In a divider using complementary staircase divider modules, this can be achieved at the expense of an additional invertor stage every other decade. An alternative method is to make all the divider stages similar, the inverting amplifiers now being placed between the divider stages.

It has been shown how a frequency divider with a variable division ratio can be produced using the staircase frequency divider as the basic module. A model of the system described functioned, with suitable values of C_1 and C_2 in each stage, up to a maximum clock frequency of about 5Mc/s. At higher frequencies than this the inherent system delay is too great causing mistriggering. Waveforms obtained from the model are given in Figs. 13 and 14 The basic divider circuit, without any form of feedback, is capable of operating from clock frequencies up as high as 10Mc/s.

Acknowledgments

The author wishes to express his thanks to his colleagues in the Applications Engineering Group, Microelectronics Division, in particular Mr. A. D. Carroll, for their assistance and encouragement and to The Director of Engineering & Research, The Marconi Company Limited, for permission to publish this article.

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A New Fourier Synthesizer Based on the Emitter Coupled Multivibrator

By A. D. Bond*, B.Sc., Ph.D.

An instrument is described which generates a fixed-frequency sinusoidal fundamental waveform and its harmonics up to the tenth. All harmonics are continuously variable in amplitude and phase, and may be summed to synthesize a variety of waveforms.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

THE electronic waveform-synthesizer described by Fox[1] is an ideal instrument for demonstration and instruction in the principles of harmonic analysis and synthesis, modulation theory and the elementary properties of signal spectra. The instrument described in this article is identical in principle with Fox's and differs only in its circuit. Its capabilities are the same, so that no applications are described here, and the precision attainable is of the same order of magnitude.

Principle of Operation

The operation of the instrument is explained by Fig. 1. A tuning-fork controlled oscillator provides a basic reference waveform, very stable in frequency. From the reference waveform are derived a pure sinusoidal waveform, the fundamental of frequency f_0, and a train of short pulses, repetitive at the fundamental frequency and used to synchronize the remainder of the system.

Each harmonic is generated by a separate oscillator, tuned to the correct frequency, and locked into synchronism with the fundamental by means of the trigger-pulse train. Continuous variation of its time-phase, relative to the fundamental, is effected by delaying the trigger pulse prior to its application to the harmonic oscillator. Every harmonic channel, therefore, has as its basic elements a linear delay unit, capable of continuous variation over the complete duration of a harmonic cycle, an auxiliary trigger-pulse generator, and a sinusoidal oscillator free-running at a pre-set frequency but capable of synchronizing with a sub-harmonic pulse train. With careful design, the auxiliary trigger-pulse stages may be omitted since the harmonic-oscillator circuit employed will synchronize from the output of the time-delay unit. However, design values are much less critical and the overall instrument is more reliable if they are included.

The synthetic waveform is produced by adding the fundamental and harmonics in a summation (operational) amplifier. This method is preferred to Fox's since the adjustment for amplitude calibration is extremely simple and involves no interaction between channels.

* *The Queen's University of Belfast.*

The Oscillators

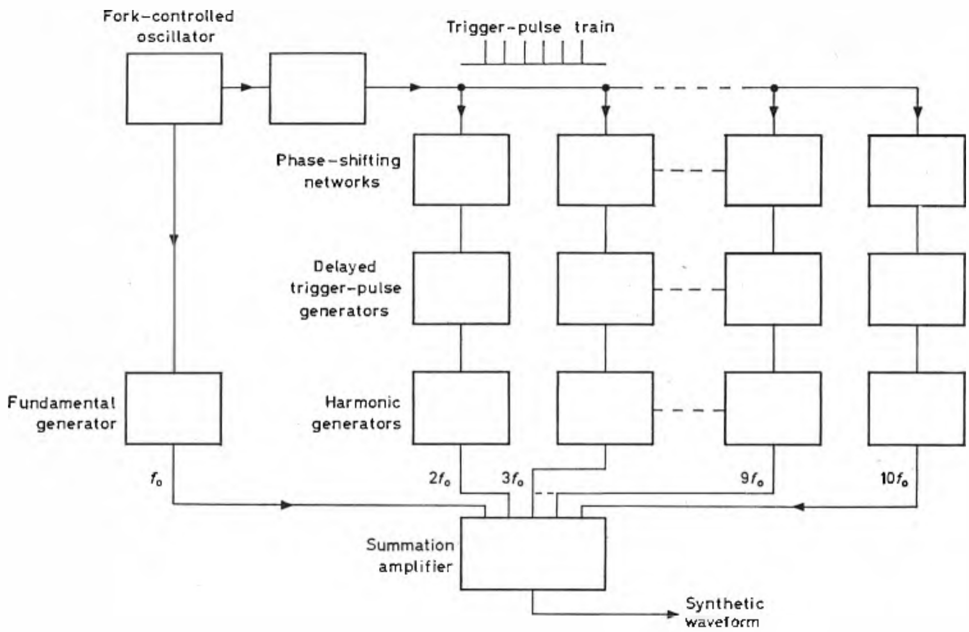
Master Oscillator and Trigger-pulse Source

The master oscillator for the system is frequency stabilized by an electromechanical circuit, for the reasons discussed by Fox. The circuit employed, Fig. 2, is effectively a free-running Schmidt trigger circuit in which oscillations are maintained by a frequency-selective positive feedback loop around the first stage. The tuning-fork used in the frequency-selective network is mounted on a solid steel base and driven by a 2kΩ headphone earpiece set on an adjustable seating approximately 2mm from the side of one prong. The earpiece coil forms the collector load for VT_1. A similar earpiece adjacent to the other prong acts as a pick-up and generates the switching waveform for the base of VT_1.

The waveform at the base of VT_1 is almost sinusoidal, with a switching impulse superimposed. It could be filtered and used directly as the fundamental, but because the harmonic distortion introduced by the switching impulse creates a rather stringent filtering problem, it is preferable to generate the fundamental separately in an additional stage.

The waveform at the collector of VT_2 is a rectangular wave with duty-cycle determined by the bias resistors at the base of VT_1. This waveform can be used, after simple differentiation and clipping, to drive the phase-shift networks. However, reliability is improved if the basic wave-

Fig. 1. Arrangement of waveform synthesizer



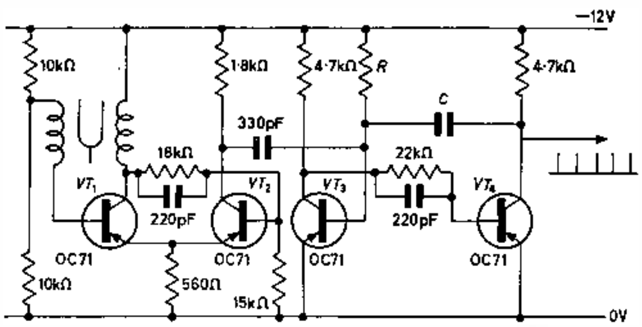


Fig. 2. Fork-controlled oscillator and trigger-pulse source

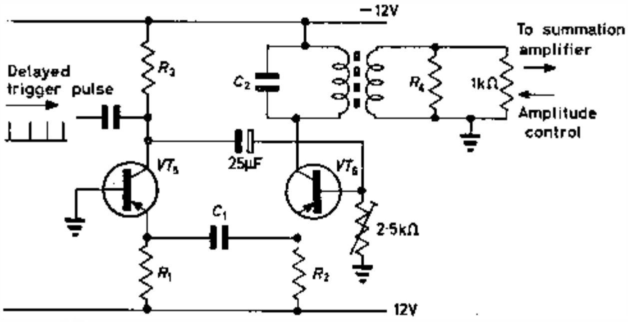


Fig. 3. Harmonic generator circuit

form source is decoupled from the rest of the system by a monostable circuit. For a fundamental frequency of 512c/s a suitable trigger-pulse duration is 50μsec, and R and C in the monostable circuit, Fig. 2, are set to give approximately this value.

Harmonic Generators

The fundamental and harmonic generators are all based on the emitter-coupled multivibrator[2,3]. The circuit has the following necessary properties.

(1) Designable to close tolerance.

(2) Easily triggered, and locks-in stably with a synchronizing waveform of sub-harmonic frequency.

(3) Stable in frequency, since the timing components are in the emitter circuit, thus eliminating frequency drift due to base current fluctuations.

(4) Free-running frequency easily adjusted by trimming component.

(5) Output collector independent of timing circuit.

The last property permits the insertion of a tuned load at the output collector VT_6, Fig. 3, so that the circuit generates a good square wave, on VT_5, and a good sine-wave, on VT_6 simultaneously. The output is transformer-coupled to the summation stage, through a linear potentiometer which is used as the harmonic amplitude control.

The detailed design of the circuit as a free-running multivibrator has been fully discussed elsewhere[2]. In the recommended procedure, the choice of operating current I_c is arbitrary, but once chosen, it determines the size of emitter resistors R_1 and R_2. The emitter capacitor C_1 is then chosen, knowing R_1, R_2, the permitted emitter voltage swing and the period of oscillation. The choice of operating current is thus limited only by the operating frequency and the sensible selection of component values, and generally speaking, the larger the current the more stable is the action of the circuit.

The introduction of a tuned circuit as the output collector load limits the choice of operating current. For best selectivity, the loaded dynamic resistance R_d of the tuned circuit should be as large as possible. However, the approximate peak-to-peak voltage swing at the collector of VT_6 is I_cR_d, and this should not exceed about 10V, otherwise the output waveform breaks through to the timing circuit. This consideration sets an upper limit on the level of operating current. The resistor R_4 is used to adjust the dynamic resistance to a suitable level, and it is found that even with reduced selectivity due to this damping resistance, the total harmonic distortion in the sinusoidal output waveform is well below 1 per cent of the maximum fundamental level. Typical component values, differing slightly from unit to unit, are $R_1 = R_2 = 5{\cdot}6\text{k}\Omega$, $C_1 = 1\mu\text{F}$, $R_4 = 150\Omega$, $C_2 = 0{\cdot}07\mu\text{F}$, for an operating current level of 2mA and frequency 3072c/s (i.e. the 6th harmonic of 512c/s). Fine frequency control is provided by the 2·5kΩ variable resistor at the base of VT_6, and the trigger-pulse may conveniently be applied to the collector of VT_6.

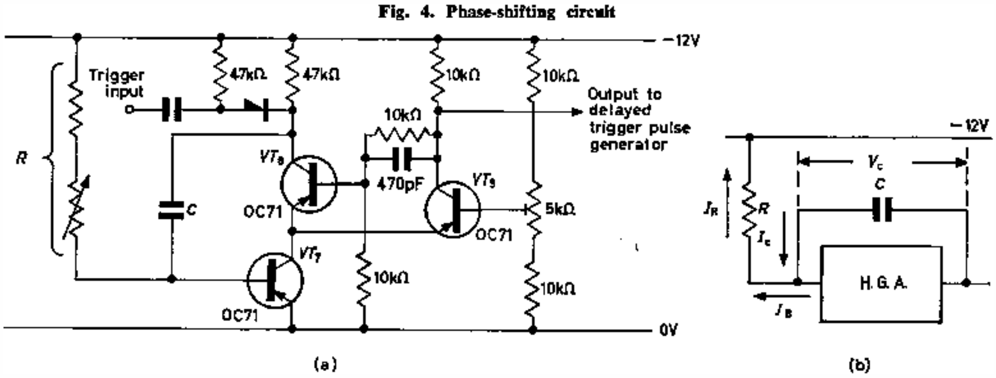
Phase-Shifters

Control of the phase-angle of each harmonic waveform, relative to the fundamental, is carried out by delaying the trigger pulse which initiates each switching cycle in the emitter-coupled multivibrator, and which thus determines the zero crossing of each harmonic cycle. The time-delay circuit[4], Fig. 4(a), is based upon the well-known Phantastron, and again is preferred for its designability, stability and ease of triggering. Delay-time is continuously variable, and is linearly related to the setting of a variable resistor.

The circuit is effectively a self-gating Miller integrator in which the components R and C determine the integrator time-constants. Transistors VT_7 and VT_8 together function as a high-gain cascode amplifier stage, and VT_9 as the switch which terminates the Miller run-down. The run-down is initiated by the basic trigger-pulse train, and causes a pulse, of controlled length, at the collector of VT_9 which is used to switch a monostable circuit and so generates a trigger-pulse, delayed with respect to the reference pulse train.

In terms of the phase angle of the harmonic waveform,

Fig. 4. Phase-shifting circuit



it is necessary to be able to delay the trigger pulse with continuous variation between plus and minus 180°, i.e. over one complete period of the harmonic. It is also preferable that these limits correspond to the end-stop settings of a variable resistor. If T_0 is the fundamental period, the delay-time of the trigger pulse for the second harmonic will vary between $(T_0/2 + T_0/4)$ and $(T_0/2 - T_0/4)$, i.e. the second harmonic basically is triggered in phase with the second half-cycle of the fundamental but may be advanced or retarded by a complete harmonic half-cycle. Similarly, the trigger-pulse delay for the n^{th} harmonic varies between $T_0(1 + 1/n)/2$ and $T_0(1 - 1/n)/2$. Thus with $f_0 = 512$c/s and $T_0 = 1{\cdot}95$msec, the sixth harmonic, for example, has a period of 325μsec. The corresponding delay generator therefore is designed to produce a pulse variable in length between 810 and 1 135μsec.

The Summation Amplifier

The summation amplifier follows normal analogue summation practice, and the circuit employed is shown in Fig. 5. The open-loop forward transfer-impedance of the basic high-gain amplifier, Fig. 5(a), is not less than 5V/μA between the frequency range 100c/s to 3kc/s. Thus in the closed-loop configuration, Fig. 5(b), the scaling and summation operations are accurate to well within 1 per cent over the operating frequency range of the instrument.

Each harmonic is applied to the summation amplifier directly from the centre-tapping of its amplitude control, and may be switched out if not required in the summation. To avoid overloading the output stage, for example when all harmonics are added in phase at maximum amplitude, a reduced sensitivity switch is provided, which short-circuits part of the feedback resistor.

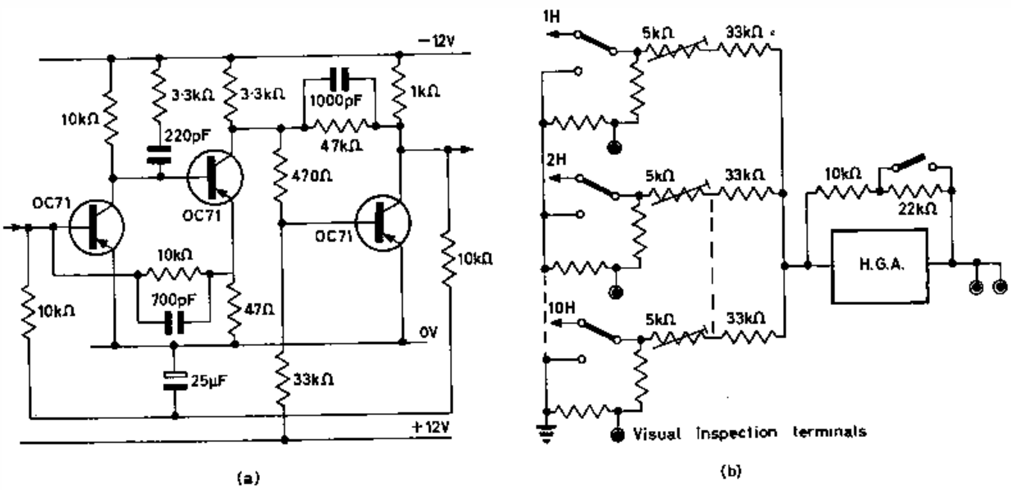


Fig. 5. The summation amplifier

In setting up the circuit it is useful to consider the junction of the timing components R and C as the virtual earth of an operational amplifier, as in Fig. 4(b). Then summing currents at the virtual earth gives $I_R = I_C + I_B$, but because the cascode operating current during the Miller run-down is only a few microamperes, the base current I_B may be neglected. With a negative supply voltage of −12V, the linear voltage sweep at the collector of VT_8 is about 10V, and this is adjustable at the base of VT_9. The current balance at the virtual earth becomes:

$$I_R = I_C$$

i.e.:

$$12/R = C(dV_c/dt)$$

$$\simeq C(10/\tau). \text{ where } \tau \text{ is the sweep-duration}$$

or:

$$R \simeq 1{\cdot}2\,\tau/C$$

Capacitor C is selected to ensure that a variable resistance of standard type will cover one harmonic period exactly. For example, for a harmonic period of 325μsec, the capacitance necessary with a 25kΩ resistor, say, is 0·016μF. When the variable resistor is fully out, the additional series resistance necessary for a delay of 810μsec is 1·2 × 810/0·016, i.e. 61kΩ. With the variable resistor fully in, the total base resistance R is then 86kΩ, resulting in a time-delay of 1 135μsec. A similar procedure is followed for each of the trigger-delay units, and this rough design method is found to give good accuracy.

Each channel is calibrated, with the amplitude of the fundamental as basis, using the 5kΩ trimming resistors in the summation amplifier input circuit. All amplitude controls are mounted in line on the front panel, each with a linear scale graduated from 0 to 1, so that the harmonics are added in as a fraction of the maximum fundamental level, rather than at an absolute level. Each phase control potentiometer is mounted beside the corresponding amplitude control, and has a linear scale graduated from −180° to +180°. Inspection terminals are provided so that each harmonic component may be displayed, as well as the synthetic waveform.

The entire instrument operates from a simple unstabilized 12V supply, and draws a total current of 250mA.

Acknowledgment

The author wishes to thank Professor J. C. West for the facilities provided, and Mr. M. W. Conroy and Mr. J. A. Nelson who carried out the development and construction work as part of a University degree project.

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A Rock-Generator Magnetometer

By P. I. Ghelfan

This article describes a direct reading rock generator magnetometer with improved sensitivity and simplified handling, for weak remanent magnetization measurements.
An electronic compensating device cancels the false magnetic effect, unassociated with the sample.
In order to avoid difficult shielding problems only RC circuits are used.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

MEASUREMENTS of the very small magnetic fields associated with the remanent magnetization of sedimentary rocks require high sensitivity magnetometers.

One of them is the rock-generator magnetometer in which a rock sample is rotated and induces in a coil a small voltage. The phase of this voltage is compared with that of a voltage of known phase obtained from the same rotating system. This comparison determines the direction of the perpendicular to the rotation axis magnetic component as against some fixed line on the rock sample. The main requirement for this measurement is to obtain a high signal-to-noise ratio.

With this end in view some rock generators use selective amplification[1,2], others both selective amplification and the selective effect of a phase sensitive detector with a large time-constant[3,4].

The sensitivity increase obtained by narrowing the bandwidth of the selective amplifiers in the first case is limited by the phase angle errors produced by the phase shift fluctuations.

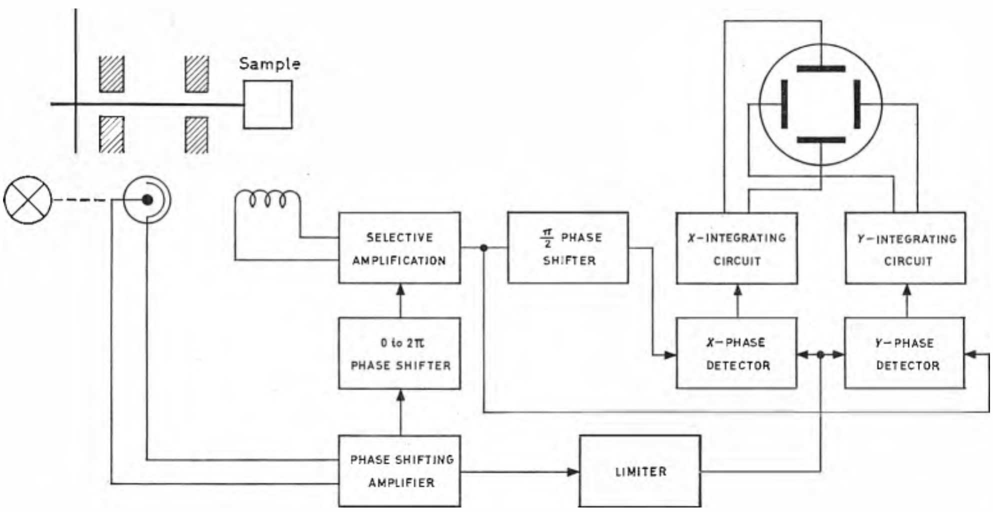

Fig. 1. Arrangement of the magnetometer

In the second case the main selective effect is given by the phase sensitive detector, with the selective amplifier only preventing overloads due to the external magnetic fields. This selective effect, and as a consequence the sensitivity, increase with the time-constant of the phase sensitive detector. However, this increase is limited by the long measuring time which results, a step-by-step method being used.

The magnetometer described in this article is based on the second variant but the step-by-step method is avoided by using a direct reading device, so that high sensitivity at a reasonable measuring time is obtained.

Possible false signals due to magnetic contamination of the rotating system are cancelled by an electronic compensation method.

Principles of Operation

The general arrangement of the magnetometer is shown in Fig. 1.

The signal induced in the coil after selective amplification, is fed to a polar presentation device which is composed of a 90° phase shifter, two phase sensitive detectors, two integrators achieving the required time-constant, and a cathode-ray tube.

The phase reference voltage is obtained by an optical system consisting of a photocell, a disk attached to the rotating system, a linear filament used as a light source, and a cylindrical lens.

An opaque portion of the otherwise transparent disk interrupts the light, thus rectangular pulses are produced on the photocell load. The photocell output is suitably amplified, limited and phase-shifted before being applied to the phase sensitive detectors.

The reference phase voltage and signal channels are nearly identical as far as phase shift and frequency characteristics are concerned. This results in only a small error in the phase determination; the false shift-frequency dependence, introduced by the signal channel selective amplifier, being partly cancelled.

A fraction of the photocell output voltage is fed through a variable 0 to 360° phase shifter to the first stage of the selective amplifier for compensating false signals produced by magnetic contaminations of the specimen holder, or the vibration of components or by any constant magnetic field with a frequency equal to that of the signal.

The frequency was chosen taking into account that the output signal-thermal noise ratio at a suitable time-constant of the phase detector is satisfactory at $f_0 =$ 30c/s and that the difference between this and the frequency of the mains is sufficient to ensure elmination of inductive effects from the latter.

Circuit Considerations

The selective amplifier (Fig. 2) is comprised of seven stages.

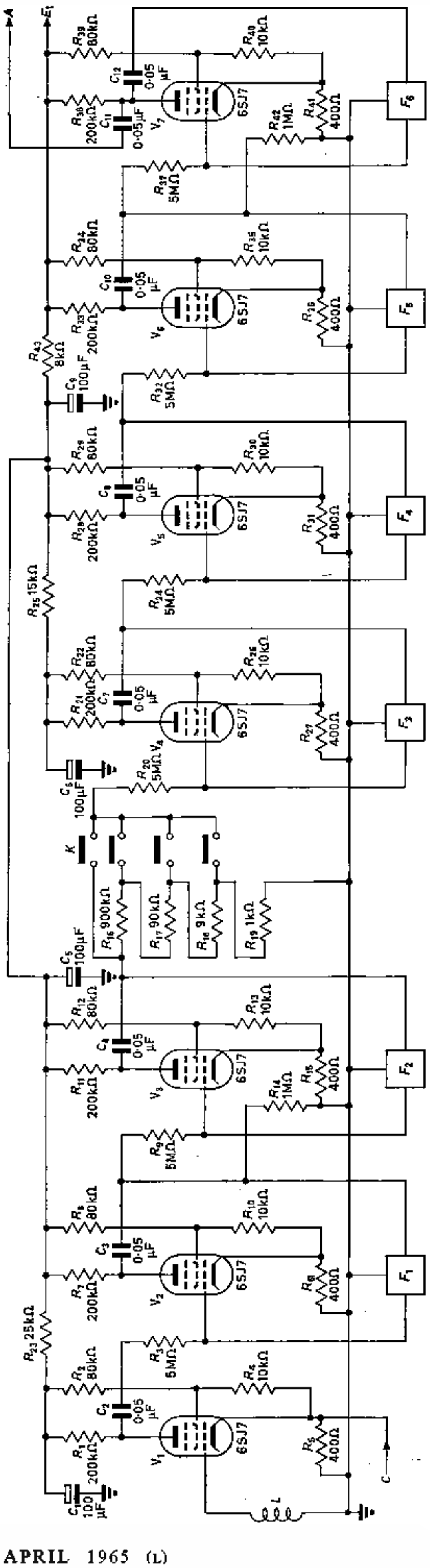

Fig. 2. Selective amplifier

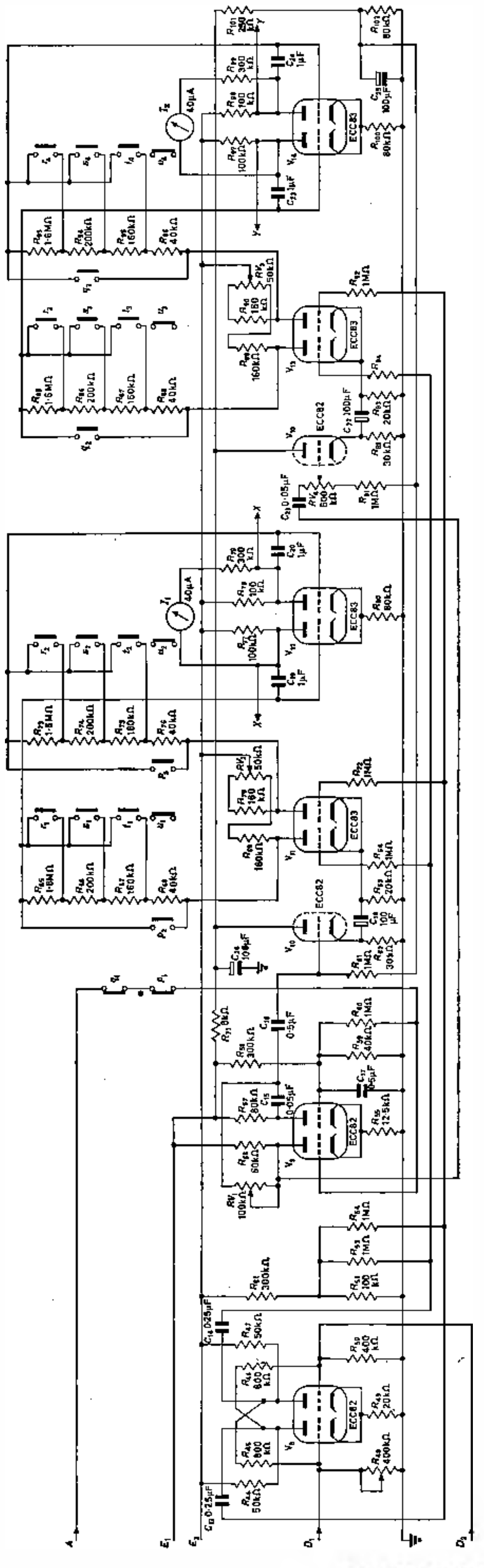

Fig. 3. Polar presentation display

The first stage is provided with a special anti-microphonic holder and is followed by six identical *RC* selective stages with twin-T networks (F_1 to F_6) tuned on 30c/s.

The amplifier has a frequency characteristic providing attenuation of the component by stray fields. The gain can be changed by switch K to $3 . 10^5$, $3 . 10^4$, $3 . 10^2$.

The anode voltage is distributed by an *RC* network (R_{43}, C_9, C_5, R_6, C_1) through which the 30c/s feedback voltage fed by the last stage to the second valve grid is phase shifted nearly 180° to improve stability.

A compensating voltage produced in the reference voltage channel is applied to the first valve cathode.

The circuit diagram of the polar presentation display is given in Fig. 3.

The reference voltage is applied in order to drive the flip-flop circuit V_9 which feeds a square wave voltage to phase detectors V_{11} and V_{13}. The amplified signal is applied to V_9 which achieves the 90° phase shift between the inputs of phase detectors.

In order to enhance the noise-to-signal ratio of the phase detectors, adequate measures were taken for their operation with small signals. Thus a high stabilization of the reference voltage amplitude and shape is achieved.

The phase detector d.c. outputs are applied to the integrating circuits composed of the switchable resistors R_{65} to R_{68}, R_{73} to R_{76}, R_{85} to R_{88}, R_{93} to R_{96} and the equivalent input capacitances of the d.c. amplifiers, V_{12} and V_{14}.

V_{12} and V_{14} amplifier outputs are applied to meters M_1 and M_2 and to the plates of a c.r.t.

The integrating circuits provide four time-constants: 2, 10, 20 and 100sec. The zero adjustment of the phase detectors is independent. (Contacts p_1, p_2, p_3 and q_1, q_2, q_3).

The reference voltage produced by the photocell is suitably amplified by the *RC* selective amplifiers with twin-T networks V_{16} and V_{17} as shown in Fig. 4, which achieves the necessary phase shift/frequency characteristic.

Thus the phase shift introduced by one stage is:

$$\phi_r = \tan^{-1}\left[-YQ_{1e}\right]$$

while that introduced by a signal channel stage is:

$$\phi_s = \tan^{-1}\left[\frac{YQ_e}{1 + Y^2Q_e^2}\right]$$

where $Y = 2\Delta f/f$.

Here Q_{1e} is the equivalent quality factor of the reference channel amplifier stage, Q_e the equivalent quality factor of the signal channel amplifier stage, and Δf the frequency deviation.

The difference between the signal channel phase shift and the reference channel phase shift will represent the frequency dependent phase angle error.

If $Q_{1e} = (3 + \alpha)Q_e$, one may approximate it by:

$$\delta\phi = 2\tan^{-1}\left[\frac{\alpha Q_e Y - 5Q_e^3Y^3}{1 + 9Q_e^3Y^3}\right]$$

Where the parameter α is to be established in order to obtain a minimum phase error between given limits of the frequency fluctuations.

Thus it was found for a 1 per cent frequency fluctuation and $Q = 10$: $\alpha = 0{\cdot}16$ the maximum phase angle error $\delta\phi_{max} = \pm 54$min.

The reference voltage fluctuations caused by the selective amplifiers require suitable limitation so as to maintain good stability of the resulting square wave voltage half period ratio; a paramount condition for the phase detector stability.

This is achieved without using inductive elements which

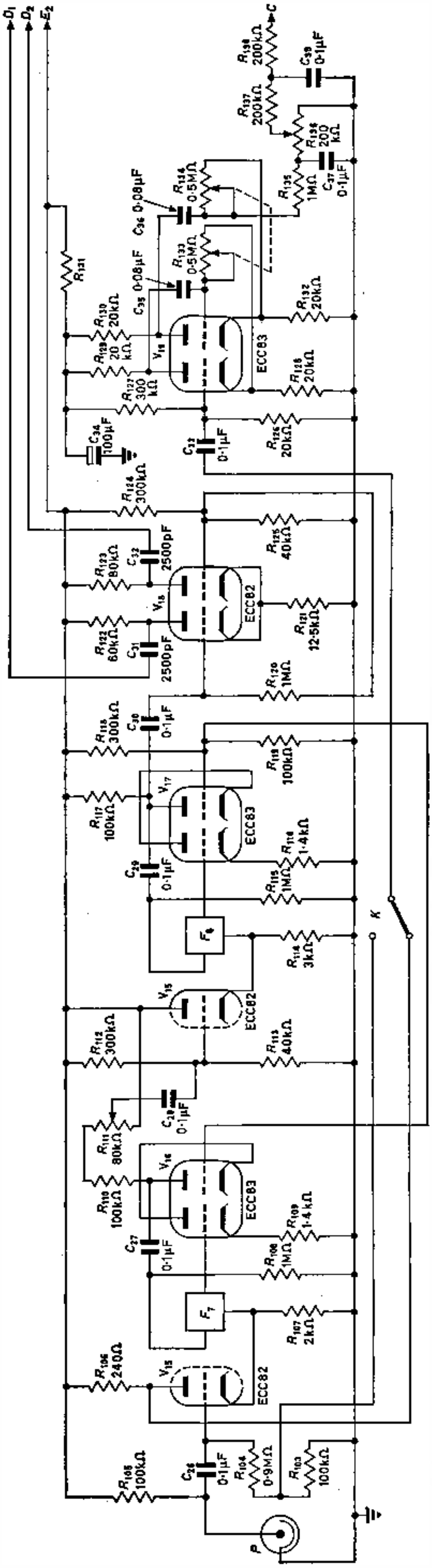

Fig. 4. Reference voltage channel

would imply difficult shielding, by the symmetric drive of a flip-flop circuit.

The required push-pull voltage is obtained with the invertor V_{18}, and applied to the flip-flop V_9 as shown in Fig. 3. The compensating voltage applied to the selective amplifiers is taken from the 0 to 2π phase shifter V_{19}, after adequate attenuation.

Conclusions

Direct reading with the help of polar presentation display increases both sensitivity and simplicity of measuring, the step by step method being avoided. At the same time false signal electronic compensation is more efficient as far as handling and extinction degree are concerned.

The calibration was made by extrapolation using $20cm^3$ specimens measured with a magnetostatic torsion apparatus.

Components exceeding 10^{-6} e.m.u./cm^3 allow angle measurements within an accuracy of 1°, and intensity measurements within 10 per cent accuracy at 1 per cent deviations of the revolving speed.

Below 10^{-6} e.m.u. (c.g.s.)/cm^3 the effect of accumulated electrostatic charges on the revolving specimen holder begins to be sensible. This effect is uncontrollable, the electrostatic charges being in continuous motion. Thus though sensitivity allows measurements even below 10^{-7} e.m.u./cm^3 they are less reliable.

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A Pneumatic Tape Transport System

A digital tape transport system, originally developed by Midwestern Instruments of Tulsa, Oklahoma is now being manufactured under licence in this country by the Control and Nucleonics Division of Plessey-U.K. Ltd.

Known as type M3000 this system employs a pneumatic pressure tape drive which, it is claimed, offers advantages over the pinch-roller and vacuum capstan systems.

A drive capstan is located on each side of the head assembly as shown in Fig. 1. The capstans rotate continuously in opposite directions and above each capstan is an air clamp, raised to permit high-speed rewind, but lowered for normal operation. The tape drive uses 15 lb/in^2 air pressure for acceleration and braking. Air at 5lb/in^2 is fed into each capstan through its centre shaft, and is forced out through the porous surface. When braked the tape is thus held away from both rotating capstans against the reservoir loop tension. The tape is braked by 15lb/in^2 air forcing it upwards against a fixed surface.

A start command diverts the air from the brake to one of the two air clamps, depending on the desired direction of tape movement. The 15lb/in^2 air pressure from the clamp collapses the separating air film between the capstan and tape. The tape acceleration is very nearly linear, with minimum peak force on the tape during acceleration.

Air supplies for forward and reverse drive and brake operation are switched through three fast acting valves which are driven by solid-state circuits. The capstan shafts are belt driven from a common motor.

Reeling System

The servo-controlled file and take-up reels are driven by d.c. motors, fed with rectified a.c. via silicon controlled rectifiers, the conduction angle being varied according to the servo torque required.

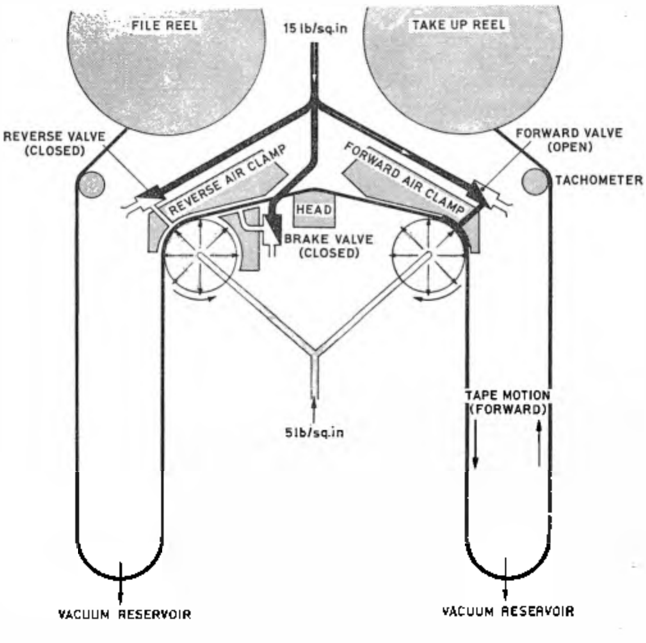
The vacuum reservoirs act as buffers between the drive capstans and the tape reels. Pneumatic transducers provide error signals proportional to the tape loop excursion in the reservoir for the reeling servos, and rate signals are derived from tachometers coupled to the two reel feed rollers. Anticipation signals are derived from the forward, reverse and stop commands.

This reeling system provides fast response with stability. Solenoid-operated reel brakes are used; these allow the reels to rotate when the solenoid coils are energized, providing braking whenever electrical power is removed.

The tape speed is up to 150in/sec with start/stop times of less than 4msec and among the advantages of this pneumatic system are the low tape wear and low skew provided by the even distribution of acceleration forces across the tape.

Plessey-U.K. Ltd is now in the process of converting the American design to use British components, to use its production methods and to convert the specification into one suitable for British computers.

Fig. 1. The tape transport pneumatic drive and tape path



A High Voltage Pulse Generator

By I. M. Paz*, M.Sc.

A high voltage pulse generator is described which delivers to a capacitive load, pulses of variable amplitude: (500 to 1 000V), width: (10 to 100μsec) and p.r.f.: (0 to 1 000p/s). The rate of voltage changes at the leading and trailing edges of the pulses are less than 1μsec/kV. The switching is performed directly on the h.v. line, the load capacitance being charged during the rise time and discharged during the fall time, by the switching elements. The instrument uses miniature hydrogen thyratrons of zero bias type as the main switching elements.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

HIGH voltage pulse generators which generate pulses of continuously variable width, p.r.f., and amplitude (in the range of several hundred volts) are valuable laboratory tools in many research fields.

They are needed in research on semiconductor and phosphorescent materials (when the performance of the above materials in various electrical fields is investigated), in controlling the deflexion of the beam in an ionic tube, etc.

Basically there are two methods of generating high voltage pulses of continuously variable width and amplitude:

The Pulse-amplifying Method

This method amplifies the low voltage pulses, produced by a conventional laboratory low voltage pulser, in a controllable amplifier.

The principles of operation are based on those of the conventional series-stabilizer of a d.c. regulated power supply translated to the pulse-techniques domain.

An instrument using this method is described by J. B. Gunn[1]. In this instrument, heavy negative feedback is used to relate the amplitude of the output pulse to a d.c. voltage derived from a precision potentiometer whose setting thus predetermines the desired pulse amplitude. The pulse amplifying method is useful when the accurate presetting of the pulse amplitude is essential.

For large pulse widths (50μsec and up) and high repetition rates (1 000p/s and up) the power dissipation of the amplifier and the high current demand from the h.v. power supply become prohibitive.

The above method is also not suitable when the output pulses must be supplied to highly capacitive loads with rise and fall times smaller than 1μsec.

The Switching Method

A brief survey of high voltage-switching techniques, used in continuously variable-width pulse-generators, is given by V. Wouk[2]. The basic circuit described there under the name "A series charge, shunt discharge circuit", is now very widely used in generators working into highly capacitive loads.

In this method, the load capacitance is charged, during the rise time of the pulse, through a 'series' switch (connected in series with the load and the h.v. power supply) and discharged at the end of the pulse, during the 'fall-time', through a 'parallel' switch (connected in 'parallel' with the load).

The 'series' switch being floating on the output, changes (during the rise and fall times) its d.c. voltage level very quickly in comparison with the usually grounded circuits that provide the control signals for it.

The generator built by Wouk[2] solves the above coupling problem by using pulse modulated r.f. for control purposes. The control signals are transferred over the different d.c. levels by transformer coupling at r.f. The modulated r.f. signals are then suitably detected producing the desired d.c. control voltages between the grid and the cathode of the main 'series' (hard valve) switches.

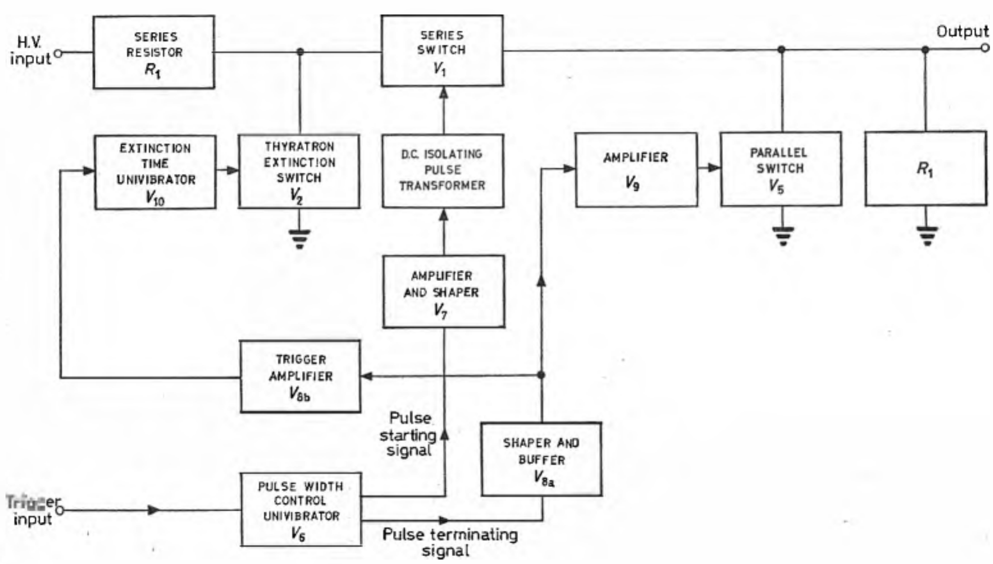
When highly capacitive loads must be driven with pulses of continuously variable width and of rise times less than 1μsec, the r.f. modulated control signal requirements become difficult to meet, making this kind of coupling impractical.

The generator described here uses the same basic 'series-charge, shunt-discharge' principle mentioned above, but introduces some novel techniques in translating this operating principle into a practical circuit.

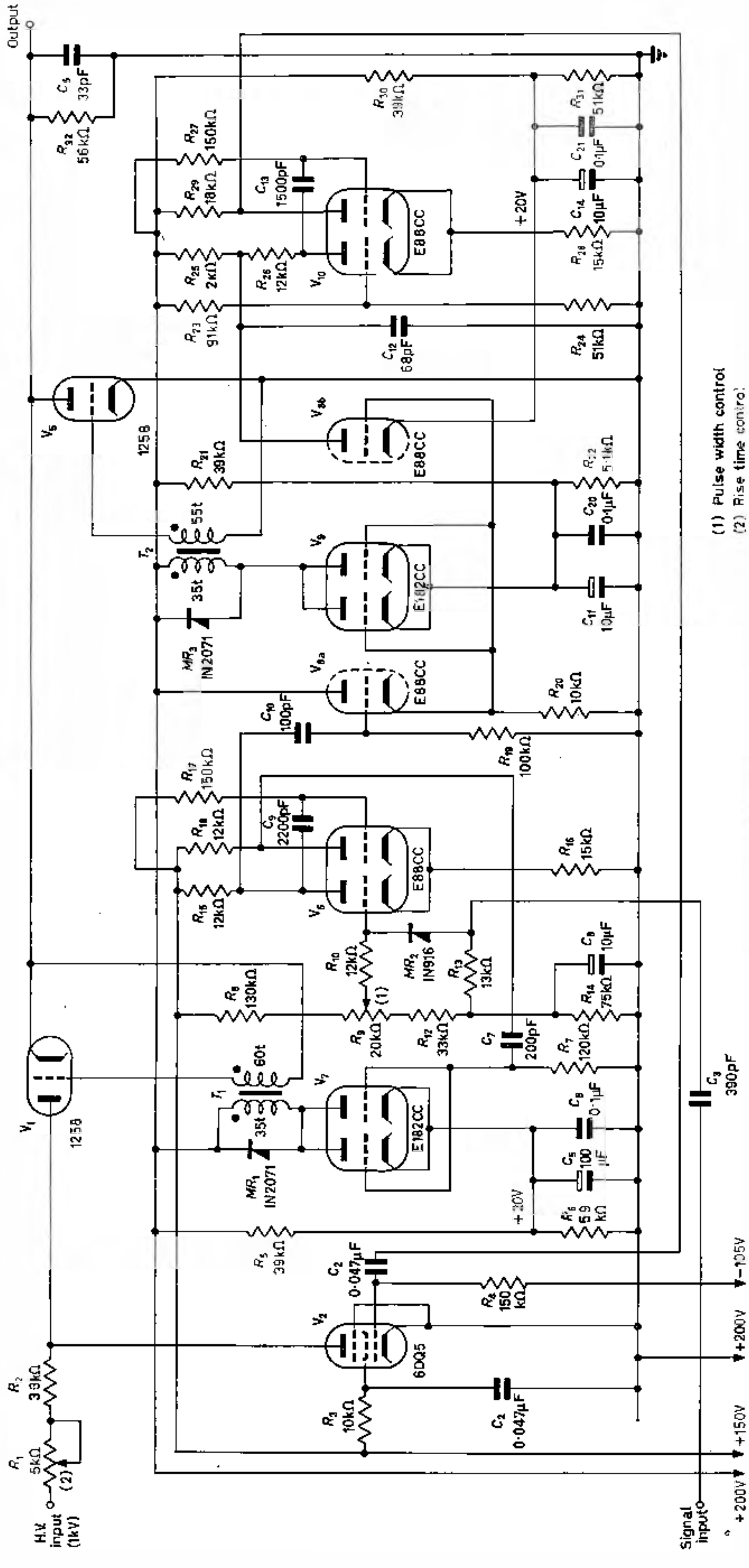
Miniature thyratrons are used as the main switches in charging and discharging the load capacitance but by the use of a particular (hard valve) extinguishing circuit for the thyratron switches, pulses of small width (1μsec) and rather high p.r.f. (1 to 2kp/s) may still be obtained.

Moreover, by using miniature hydrogen thyratrons of

Fig. 1. Principle of operation of the high voltage pulse generator



* *Haile Selassie I University, Ethiopia.*



(1) Pulse width control
(2) Rise time control

Fig. 2. The complete circuit

T_1, T_2 = pulse transformers in ferrite cup cores (number of turns as indicated)

the zero-bias type, simple pulse transformers may be used as the coupling link between the control circuits and the 'series' thyratron switch they drive. No floating circuits (other than the 'series' switch) being needed, the stray capacitances of the output line are much reduced, simplifying the circuit of the instrument.

A compact, versatile pulse generator is therefore obtained, producing pulses over a wide range of specifications with a comparatively small piece of instrumentation.

The Principles of Operation

The block diagram of the instrument, given in Fig. 1, will be used in describing its principles of operation.

The output pulse is obtained as follows:

(a) The load capacitance (connected to the output terminals of the instrument) is charged—during the rise-time—from the external h.v. power supply, through the series resistor (R_1) and the closed series-switch (V_1).

(b) The 'parallel-switch' (V_5) short-circuits, during the fall-time, the output to ground, by this, terminating the output pulse.

(c) The 'series-switch' V_1 is held closed during the whole 'pulse-width-time', allowing a constant current to flow into the resistive-load (R_L) which is connected in parallel with the output.

The voltage across the load thereby remains constant over the whole pulse width ensuring by this the flatness of the 'pulse-top'.

(d) The control signals for the two main switches V_1 and V_5 and for the thyratrons-extinction-switch V_2 are initiated in the pulse-width-control - univibrator (V_8). For each external trigger entering the 'trigger-in' terminal of the instrument, a pair of pulses is produced by the above univibrator. One of these pulses starts the output-pulse producing process and the other one is terminating it. The time interval between the above two pulses is controlled by the V_6 univibrator which thus controls the width of the output pulse.

(e) Together with the parallel-switch (V_5) the extinction-time univibrator (V_{10}) is also operated. This univibrator produces an output-pulse of a width which is prescribed by the maximum time required for safely extinguishing the main thyratron switches V_1 and V_5. The output of the extinction-time univibrator controls the thyratrons extinction switch, short-circuiting it to ground for the whole extinction period.

The reopening of the main switches is by this ensured a certain time (corresponding to the width of the V_{10} univibrator output pulse) after the termination of each main output pulse. The maximum p.r.f. of the main output pulses is therefore limited mainly by the above extinction period and the desired width of the main output pulse.

(f) The p.r.f. of the pulses is externally controlled by a standard pulse generator which feeds the 'trigger-in' terminal of the instrument.

(g) The amplitude of the main output pulses is controlled by the voltage level supplied from an external high voltage power supply connected to the h.v. in terminal.

(h) To protect the external h.v. power supply from being short-circuited (to ground) during the operation of V_5 or V_2, a series resistor R_1 is used.

By making a part of the above resistor variable, a rise time control is obtained which enables a certain overshoot-tise time trade.

The Circuit Description

The circuit diagram of the instrument is given in Fig. 2 and the waveforms that develop at various points of it, are given in Fig. 3.

The main switches (V_1 and V_5) are miniature hydrogen thyratrons, type 1258, of 'Tungsol' make. (See Appendix for power dissipation considerations).

A pulse forming cycle starts every time a proper triggering signal is applied to the 'signal in' terminal. The external triggering source controls therefore the p.r.f. of the output.

The shape of a typical triggering pulse as needed at the 'signal in' terminal is given in Fig. 3(a). This triggering signal operates the pulse width control univibrator V_6.

The potentiometer R_9 controls the widths of the output pulses obtained at the anodes of the univibrator. The width of the above pulses being used to control the width of the main output pulses, R_9 will provide the pulse width control element.

The positive pulse produced by this univibrator at its V_{6b} anode is differentiated and amplified by the peaking amplifier V_7 (anode waveform Fig. 3(g)).

The trigger obtained from the leading edge of the univibrator pulse is transformer-coupled through T_1 to the grid of the series thyratron switch V_1, firing it, and thus starting the leading edge of the main output pulse.

The firing of the V_1 thyratron switch enables a heavy current to start flowing through the series resistors (R_1, R_2), the now-conducting thyratron (V_1) and the load capacitance.

The load capacitance will therefore start charging, towards a voltage level, mainly determined by the external h.v. power supply and the potential divider, R_1, R_2, R_{32}

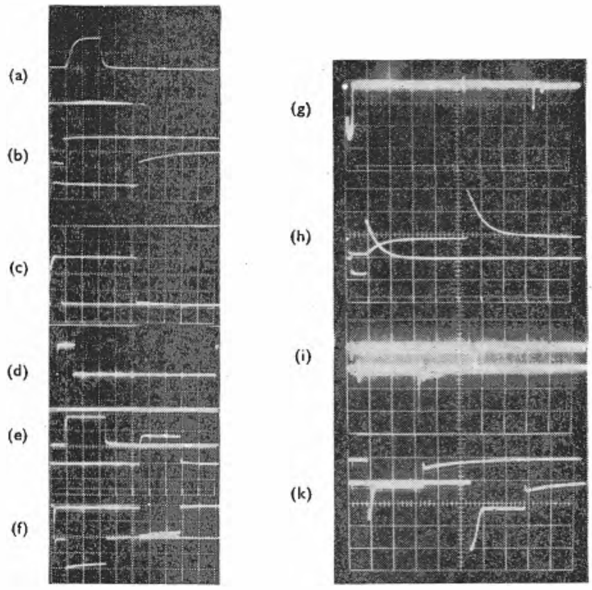


Fig. 3. Waveforms

(a) *Typical trigger signal (min. amplitude, min. width, max. rise-time) as needed at the 'signal-in' terminal. Scale: 1µsec/cm, 20V/cm*

(b) *Pulse available at the anode of V_{6a} (max. and min. width). Scale: 20µsec/cm, 20V/cm*

(c) *I_D at the anode of V_{6b}*

(d) *Output pulse. Scale: 500V/cm, 100µsec/cm*

(e) *Anode of V_{10b}. Scale: 20µsec/cm, 50V/cm*

(f) *Anode of V_2 (max. and min. pulse width at the output). Scale: 20µsec/cm, 500V/cm*

(g) *Anode of V_7. Scale: 10µsec/cm, 10V/cm*

(h) *The output obtained at the cathode of V_{8a}. Scale: 20µsec/cm, 10V/cm*

(i) *Triggers obtained at the anodes of V_9. Scales: (min. out pulse width) 5µsec/cm, 20V/cm (max. out pulse width) 20µsec/cm*

(k) *Pulse at anode of V_{10a} (at max. and min. out pulse widths). Scale: 20µsec/cm, 50V/cm*

and with a time-constant determined by the load capacitance itself and the series resistors R_1, R_2.

This charging time-constant of the output capacitance determines also the rise time of the main output pulse leading edge. A rise time control is made available, by making the series resistor R_1 variable. This rise time control enables a rise time overshoot trade, when a monotonic output pulse waveshape is needed.

When the charge delivered to the output capacitance brings the voltage across it to 90 per cent of the desired output amplitude, the rise time period is over but the series thyratron switch continues its conduction, because of the R_1, R_2, R_{32} potential divider which permits a small current (just enough to maintain ionization of the tube) to flow through it.

A 'trickle-charge' current, which adjusts itself to overcome the discharge of the load capacitance during the pulse time, flows therefore all the time into this capacitance, so ensuring the flatness of the pulse top.

To terminate the main output pulse, the load capacitance has now to be quickly discharged by short-circuiting it to ground through the paralled thyratron switch V_5. The pulse obtained at the anode V_{6a} of the pulse width univibrator (waveform Fig. 3(b)) is used for this purpose.

The trigger, obtained by differentiation of the trailing-edge of the above pulse, fires now (through the V_9 amplifier) the parallel switch V_5.

The pulse-forming period is now completed but the two main thyratron switches are still conducting and a means must now be provided to extinguish them, in order to prepare the instrument for a new pulse-forming period.

For this purpose the pentode V_2 (a 6DQ5) is provided, that short-circuits the anode of V_1 to ground for a period of 50μsec (the extinction period) extinguishing the two thyratron switches V_1 and V_5. The extinction period being longer than the longest deionization time of the thyratron switches their reliable extinction is ensured. To control the V_2 thyratrons extinction switch the extinction time univibrator (V_{10}) is used.

This univibrator produces at its V_{10b} anode a 50μsec pulse, (waveform Fig. 3(e)) of an amplitude large enough to bring the V_2 thyratrons extinction switch into heavy saturation.

The triggering of this univibrator is done through the (buffer) trigger-amplifier V_{8b} by the same waveform which operates the V_9 amplifier and therefore in synchronization with the fall time of the main output pulse.

Results

The specification obtained from the laboratory model of the instrument is as follows:

(a) *Amplitude and Width* of the output pulse (waveform Fig. 3(d)) are variable in the ranges 500 to 1 000V and 10 to 110μsec respectively, the amplitude being controlled by the variable external high voltage power supply.

(b) *Triggering*: the instrument has to be triggered with positive pulses from a conventional (external) pulse generator.

(c) *The P.R.F.* may be varied in the range 0 to 1 000p/s. A higher p.r.f. is possible (3 to 4 fold) when a higher current demand from the external high voltage power supply can be tolerated. The average current demand may be calculated, considering that after every pulse of maximum amplitude, a 320mA pulse of current is needed for a period of 50μsec (through the thyratrons-extinction switch).

(d) *The Amplitude rate of change* (in the rise as well as the fall time) is faster than 1kV/μsec. The rise time, as already mentioned, may be changed using the rise time control potentiometer, which provides also a means of overshoot attenuation.

(e) *Load*: The instrument described is able to feed a capacitive load of up to 100pF.

Conclusions

By using miniature zero-bias thyratron switches in a 'series charge, shunt discharge' circuit, clean rectangular pulses of variable amplitude, width and p.r.f. can be generated into a capacitive load.

The p.r.f. of the so generated pulses may be increased almost up to the limits imposed by the maximum required pulse width and the maximum deionization time of the thyratron switches, provided a suitable extinguishing circuit is used.

When designing the mechanical lay-out and the wiring of the instrument, much care must be taken to minimize stray capacitances, shorter rise times can then be obtained with switches of smaller power dissipation, leading to an instrument of smaller external dimensions.

Appendix

The power dissipation considerations which permit the use of miniature thyratrons as the main switches are as follows:

To charge the almost purely capacitive load, the series switch has to supply most of its current in peaks of several amperes during the rise time of each pulse only.

Supposing a constant current source is available, for a 1μsec rise time, amplitude of 1 000V and a load of 100pF, a current of 10A will be needed for the whole rise-time period.

Considering now a maximum p.r.f. of 1 000p/s the duty ratio will be less than 0·001.

The voltage drop across the thyratron switch during conduction being of the order of 20V, the average power dissipation that this switch must withstand will have to be of the order of 0·2W.

Miniature thyratrons will therefore be suitable.

The parallel switch will have to withstand a similar power dissipation, having to conduct only during the fall time of the output pulse (in order to discharge the load capacitance) which is also of the order of 1μsec.

From the above consideration it is also clear that the average current taken from the h.v. power supply, through the series switch, to the load is negligibly small and the main current consumption from the above supply will be therefore the one taken by the thyratron extinction switch.

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Helicopter Navigation Systems

The development of helicopters into all-weather aircraft from vehicles largely suitable for daylight operation in visual conditions has, until recently, out-stripped progress in automatic navigation equipment. Now, Computing Devices Co. Ltd has introduced an integrated system of navigational aids which already have shown themselves to be acceptable for all forms of helicopter operations. These range from the simplest local or route flying patterns to the complication of ship-borne anti-submarine warfare. Based on simple, modular, and operationally proven units the equipment can be fitted to any type of helicopter and is offered in a variety of configurations to match any choice of sensors.

In the past, navigational systems have been unequal to rapid changes in helicopter roles or tactics. The new Computing Devices system—designated PHI-10B (Position and Homing Indicator Type 10B)—is designed to handle all current needs and changes in the foreseeable future. PHI-10B can be applied to every form of civil helicopter operation especially in areas of minimum ground aids such as oil drilling rigs, geological and forest survey, emergency operations and power transmission line surveys.

It has been operationally tested by military units and found suitable for both present and future demands. PHI-10B equipment available now, provides an integrated, helicopter navigation computing and display system. Further development for various military helicopter projects is proceeding and the design of the present PHI units will permit replacement by next generation modules as their development is completed and without loss of system effectiveness. This enables helicopters now under development to be evaluated and delivered as a complete weapons system with maximum present system performance plus economic growth potential if required.

Three units only make up the basic PHI-10B and these may be used for the routine task of en-route flying between up to 12 pre-planned fix points or destinations. More operational flexibility is available by adding only those extra units which are necessary to cover new requirements. One of the most useful of these units is, for example, a Vector Adder which allows the pilot to programme new unplanned destinations while airborne.

Operational helicopter experience has shown that the advantages accruing from a largely proven modular system are considerable. The system offered combines light weight, small size, high precision, reliable performance and easy operation. Any arrangement of the system gives full navigational information for lift-off, transit, hover approach, hover and landing.

Improved S.C.R. Firing Circuits

By D. Ciscato* and L. Mariani*, E.E.

The rapidly increasing application of s.c.r. in power control and conversion has involved the design and implementation of a large number of firing devices.

In this article two such all solid state circuits are presented, one suitable for single-phase inductive loads and the other for three-phase controlled bridges.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

Unijunction Transistor Firing Circuit for S.C.R. Single-Phase Circuit with Inductive Load

UNIJUNCTION transistor circuits are very widely used in single-phase resistive load s.c.r. firing because of their low cost, simple circuits and low power requirements. As known, highly inductive loads demand higher energy firing pulses for reliable s.c.r. operation. In this case it may be convenient either to maintain the s.c.r. gate energized throughout the conducting half-cycle or, for a lower gate dissipation, to supply it with a train of closely spaced pulses to the end of the half-cycle[1,2,3,4].

Suitable modifications of the basic synchronized relaxation oscillator in Fig. 1 make it possible to obtain, at the output terminal, a constant rate pulse train with the instant of initiation depending on a d.c. control voltage applied to the driver input. The improved circuit is shown in Fig. 2.

The pulse repetition rate depends only on the time-constant $T_1 = R_1C_1$ which may be small, while the initial delay, related to time-constant $T_2 = R_1(C_1+C_2)$ is controlled by varying the V_M voltage according to V_c. The supply voltage V_s is obtained from an a.c. line by Zener diode half-wave rectification and squaring, so the circuit is synchronized with the s.c.r. forward voltage.

Starting from t_0, Fig. 3, after a short delay due to the rise of V_M up to V_N and the charging time of C_1 ($C_1 \ll C_2$), charge of C_1 and C_2 begins through R_1 with a voltage difference $V_s - V_M$. As soon as V_A reaches the unijunction transistor peak voltage V_P, C_1 discharges through R giving the first output pulse.

Capacitor C_2 discharge is blocked by diode MR_3 so that following pulses depend only on the charge of C_1 and their repetition rate is about:

$$f \simeq \frac{1}{R_1C_1 \ln\left(1/(1-\eta)\right)}$$

which is not related with control voltage V_c.

At the end of the half-cycle capacitor C_1 begins its discharge through R_2, MR_2, MR_4, while MR_1 de-couples C_2 from the driver circuit.

Instead of unijunction transistors one may use some other device with negative resistance characteristics such as a Shockley diode, Transwitch, pnpn transistor or a pair of pnp and npn transistors. It is possible to stop the pulse train as soon as the s.c.r. fires by choosing for V_{sc} the voltage across the s.c.r.; in this case gate dissipation is lowered.

Three-Phase S.C.R. Controlled Bridge Firing Circuit

In order to regulate the mean output voltage from a three-phase controlled bridge (with three or six s.c.r.) it is necessary to apply to the gate of different s.c.r.'s 120° spaced single pulses or 60° spaced pairs respectively, with phase variable, from 0° to 180° according to a d.c. control voltage. Most s.c.r. applications in three-phase controlled bridges use separate firing circuits, one for each phase, driven by the same d.c. control voltage; the difficulty of obtaining very close matched characteristics in different units very often causes a non-uniform current sharing. For this reason some firing units with a single variable pulse displacement device were developed[5].

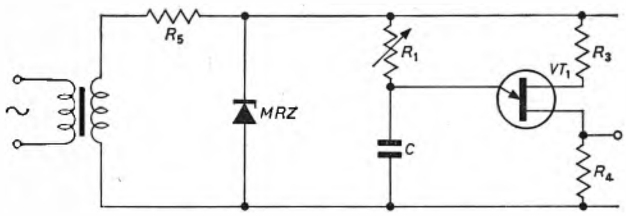


Fig. 1. Basic unijunction firing circuit

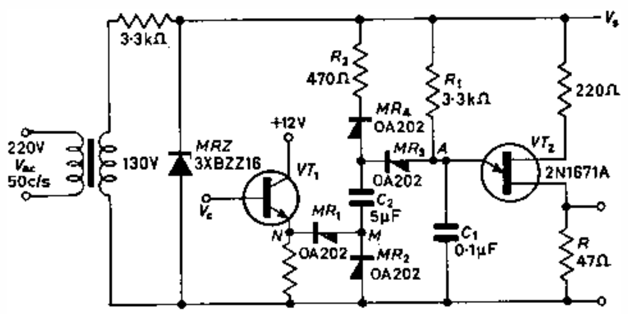
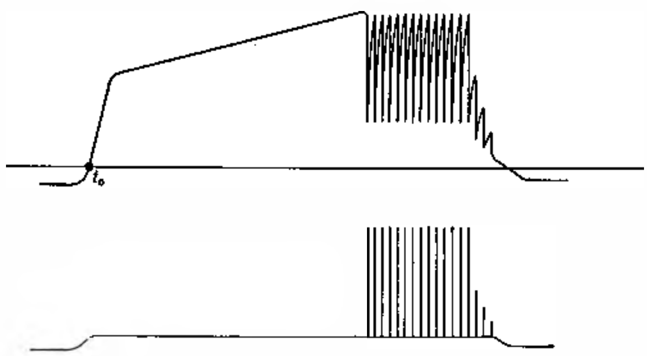


Fig. 2. Modified unijunction firing circuit

Fig. 3. Point A and output waveforms of Fig. 2 circuit



* *Electrical Institute, University of Padua (Italy).*

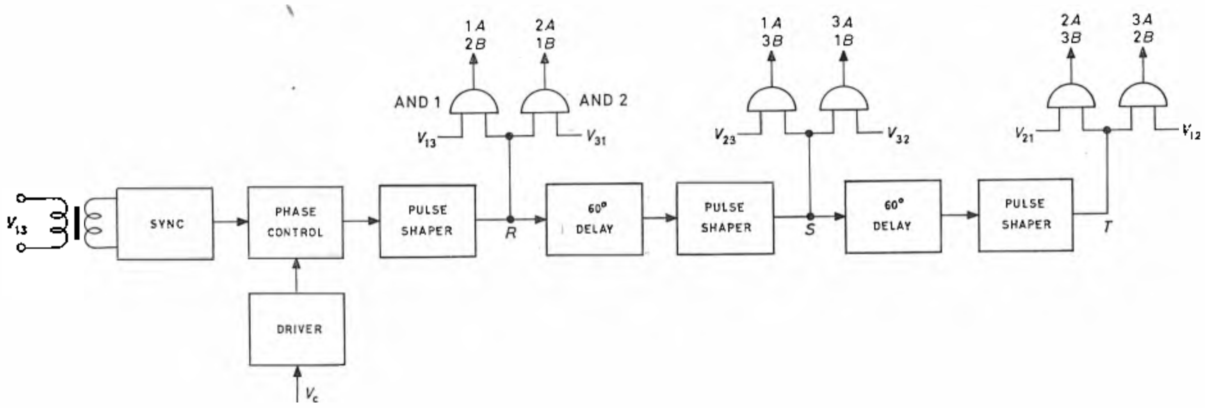


Fig. 4. Firing circuit arrangement for Fig. 5 bridge

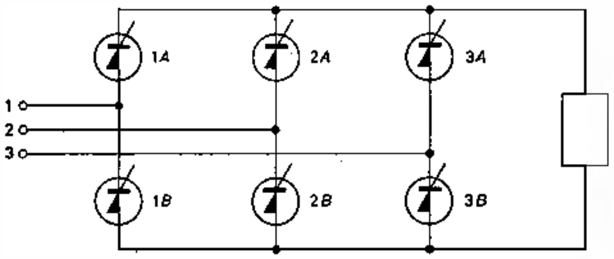


Fig. 5. Three-phase, six s.c.r. bridge rectifier

The block diagram of a simple and versatile circuit of this type is shown in Fig. 4: it is suitable for six s.c.r. three-phase bridges (Fig. 5). In this case the firing circuit must supply, every 60° electrical degrees, a pulse simultaneously to two s.c.r., one in the upper and one in the lower row, according to Table 1:

TABLE 1

0°	60°	120°	180°	240°	300°
1*A*	1*A*	2*A*	2*A*	3*A*	3*A*
2*B*	3*B*	3*B*	1*B*	1*B*	2*B*

A synchronizing pulse occurring at zero crossing of one of the line voltages, V_{13}, for instance, triggers a sawtooth with duration depending upon control voltage.

The pulse shaper provides a pulse at the end of the sawtooth, suitable for the gate drive required. The correct pulse for feeding to the various s.c.r.'s may be derived

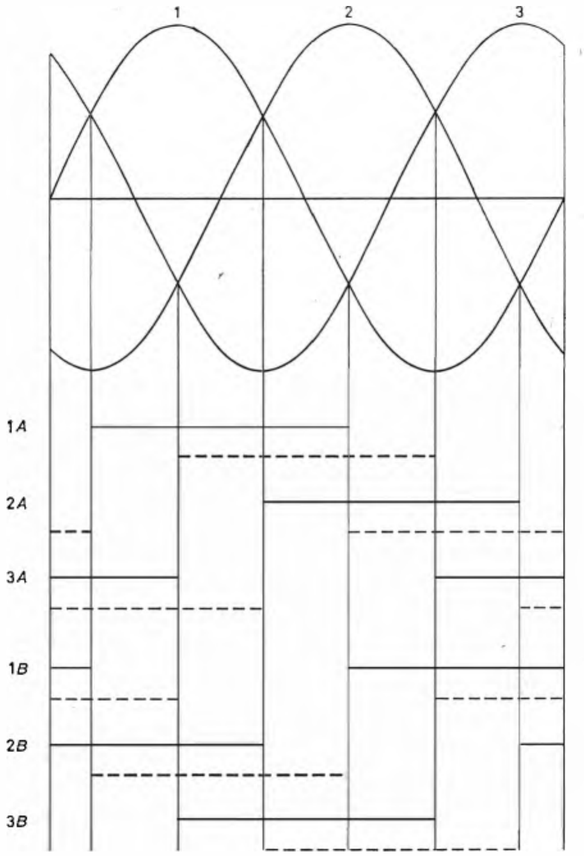


Fig. 6. Time intervals in which first pulse (full line) and second pulse (dotted line) must be supplied for correct s.c.r. firing of Fig. 5

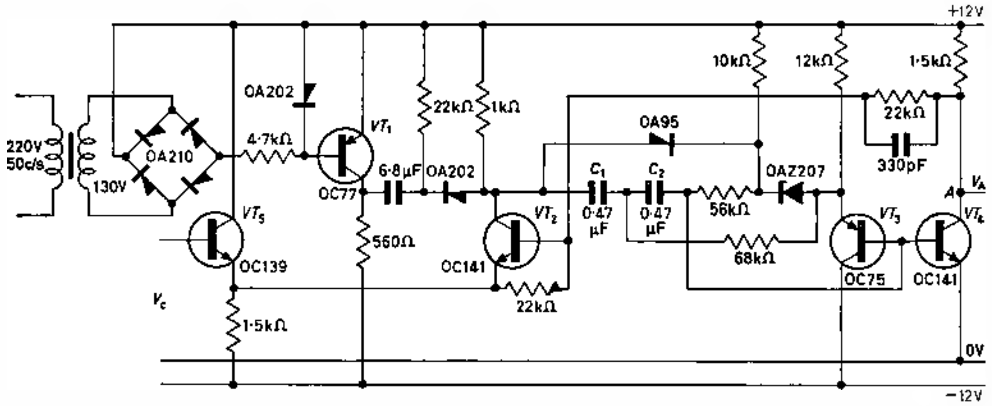


Fig. 7. Synchronized phase control

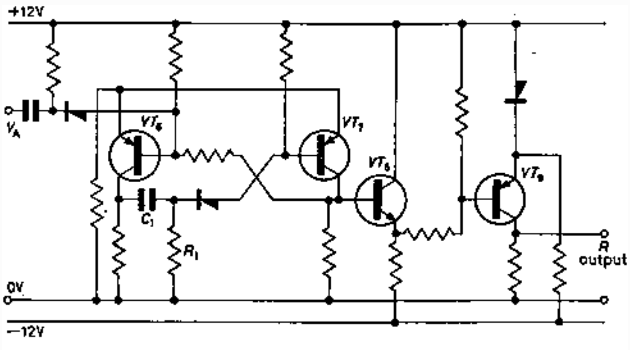


Fig. 8. Pulse shaper

from line voltages as shown from Fig. 4 and Fig. 6. The available pulses at point R are suitable for s.c.r. $1A$ and $2B$ during the V_{13} positive half-cycle, and for $2A$ and $1B$ during the negative one. A fixed delay circuit provides the first 60° phase angle delay and the second pulse shaper

VT_6 is 'off' so that the voltage at R is about 0V. When VT_6 and VT_7 multivibrator is triggered its output is a 12V amplitude rectangular pulse with a duration depending on timing circuit C_1 and R_1; this value is chosen according to the gate requirements. This rectangular pulse through VT_8 is fed by VT_9 to output R.

Gating Circuit

Fig. 9 shows the complete gating network necessary for pulse distribution according to Table 1. The rectangular pulse at R may be applied to winding 1 only if the switch VT_{10} is closed i.e. during the time interval when V_{13} is positive (Fig. 6) with winding sense of Fig. 8. During this time interval VT_{11} is 'off' so no pulse is applied to winding 1′.

The contrary case occurs when V_{13} is negative so winding 1′ is pulsed. R_1 are base current limiting resistors, and MR_1 diodes avoid voltage spiking in transistor collectors. Transformer T_1 with VT_{10} and VT_{11} implements the two AND 1, AND 2 gates of Fig. 4. Diodes MR_3 eliminate nega-

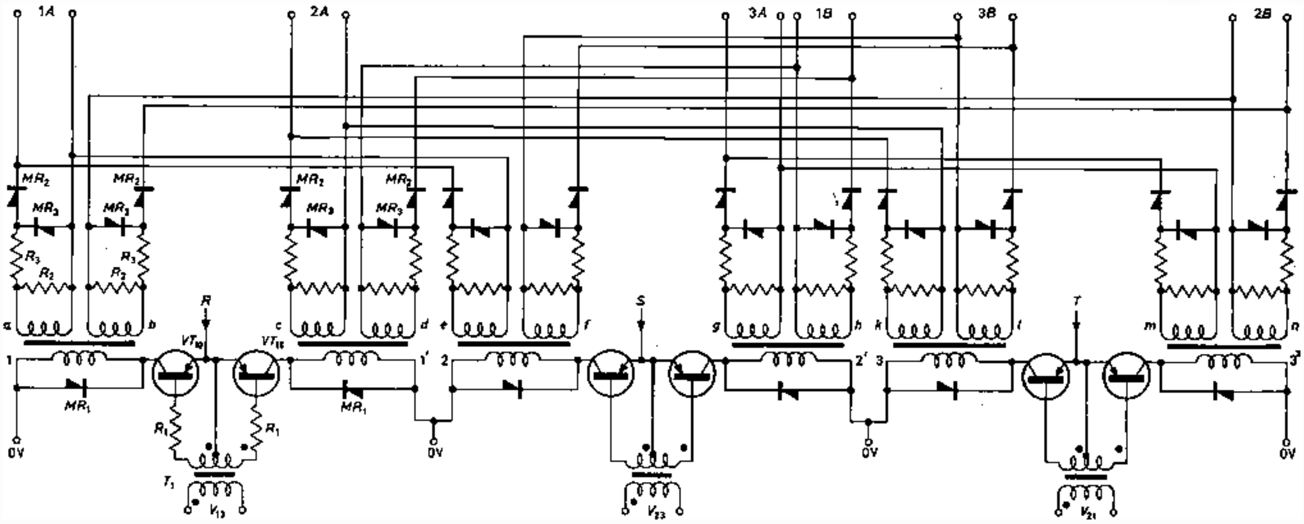


Fig. 9. Gating network for pulses distribution

action is the same as mentioned above. Pulses at point S are now suitable for $1A$ and $3B$ s.c.r. firing during the V_{23} positive half-cycle and for $3A$ and $1B$ in the negative one, likewise at point T for $2A$, $3B$ and $3A$, $2B$.

Synchronization and Phase Control

In Fig. 7 VT_1 provides a trigger pulse for each zero crossing of V_{13} line voltage, to a linear delay circuit. The one considered in this article was described by Rakovich[6]. VT_2 and VT_4 form a collector coupled monostable multivibrator with VT_4 'on' and VT_2 'off' in the stable state.

The discharge of timing capacitors C_1 and C_2 is linearized by the Bootstrap action of transistor VT_3 keeping discharge current constant.

If VT_2 is allowed to be saturated, the positive square wave duration at point A is directly proportional to the d.c. control voltage V_c at the emitter of VT_5. This circuit is temperature compensated and allows a large delay range.

Pulse Shaper

The pulse shaper (Fig. 8) is formed by a monostable multivibrator VT_6 and VT_7 triggered by the trailing edge of voltage V_A, followed by an emitter-follower VT_8 and an amplifier stage VT_9. In the stable state VT_7 is 'on' and

tive spikes, and R_3 determines the load line in the gate circuit and is chosen according to gate requirements[7].

As shown in Table 1 all s.c.r.'s are fired by two pulses 60° delayed from each other. For instance the first pulse suitable for s.c.r. $1A$ firing may be derived from winding a, and the second one from e. Diodes MR_2 perform an OR logic operation on these positive pulses producing in $1A$ output the desired gate drives.

60° Delay Circuit

This is a simple one-shot multivibrator circuit; provisions must be made for limiting temperature effects on square wave duration (diode techniques, etc.).

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A Transistorized Analogue Divider

By G. Gooberman*, Ph.D., and B. Jones*

The design of a transistorized analogue divider is described. The divider is capable of giving an output predictable to better than 0·5 per cent and linearity better than ±0·2 per cent for voltages between 0·7 and 20V. A variable mark-to-space pulse technique is used in which a tunnel diode forms the heart of the pulse generator.

(Voir page 281 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 288)

DURING the development of an analogue simulator for the steady state flow of fluids in pipes the need arose for a fairly cheap analogue divider to work over a narrow range of voltages confined to one quadrant and with an accuracy in the region of 0·5 to 1 per cent. Various types were considered but the final choice fell upon a variable mark-to-space pulse system employing a tunnel diode pulse generator so designed that its accuracy is reasonably independent of moderate changes in pulse repetition frequency. The system does not require the generation of an accurately linear ramp waveform, in fact an exponential is used. Its disadvantage is that because of the simple type of filter used the divider is only suitable for use with signals with frequencies below 1c/s.

Principle of Operation

The function of the divider is to generate a voltage V_o proportional to V_n/V_d. A block diagram is given in Fig. 1. The denominator voltage V_d is chopped by chopper A, driven from a variable mark-to-space pulse generator and the resultant negative going rectangular waveform is applied to one side of the comparator after being smoothed by filter A. A fixed voltage, V_f, is applied to the other comparator input whose output, proportional to the difference between its input voltages, controls the mark-to-space ratio of the pulses from the pulse generator in such

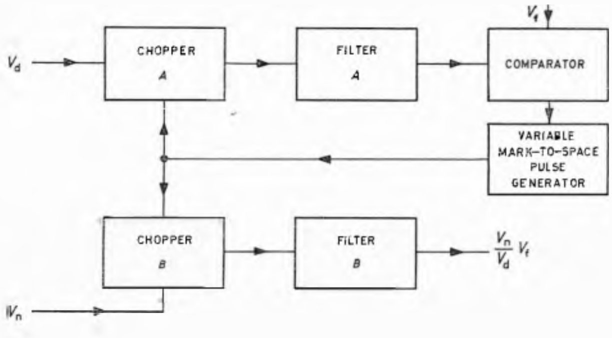


Fig. 1. Arrangement of divider

a way that the output of filter A remains constant independent of V_d. If the output of filter A were to be exactly equal to the average value of the rectangular waveform from chopper A the mark-to-space ratio of the pulse generator output would be exactly inversely proportional to V_d. The drive to chopper A is also applied to chopper B which chops the numerator voltage V_n. The resultant rectangular waveform is smoothed by filter B whose output will be proportional to V_n/V_d. In practice, the filter outputs are not exactly equal to the average value of the input waveform but by making both filters identical the errors which would otherwise occur can be eliminated.

Theory

The chopper and filter must be considered together as shown in Fig. 2. A voltage V_i is applied to the chopper input and the resultant rectangular waveform appearing at X is smoothed by R_2 and C. If the chopper is open for

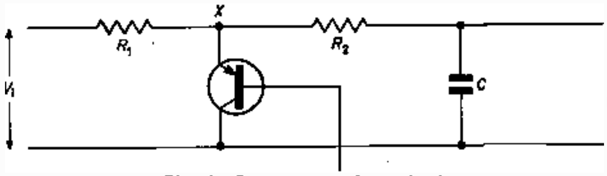


Fig. 2. Chopper and filter circuit

a time t_1 and closed for a time $T - t_1$ and provided that the filter time-constant is long compared with T, the steady voltage V_c appearing across the capacitor is given by:

$$V_c = \frac{V_i t_1}{T(R_c/R_d) + t_1[1 - (R_c/R_d)]} \qquad \ldots\ldots (1)$$

in which $R_c = R_1 + R_2$, the resistance through which C is charged and $R_d = R_2$, the resistance through which C discharges. The comparator and variable mark-to-space pulse generator control t_1 such that the output of filter A equals V_f which is a constant. Thus after rearranging equation (1) and replacing V_i by V_d and V_c by V_f:

$$t_1 = \frac{V_f T}{V_d(R_d/R_c) + V_f[1 - (R_d/R_c)]} \qquad \ldots\ldots (2)$$

It will be noticed that t_1 is exactly inversely proportional to V_d only when $R_c = R_d$.

One can now find the output voltage of filter B, V_o, when chopper B chops V_n by inserting equation (2) back into equation (1) in which R_c and R_d are replaced by R_c' and R_d', the resistances appertaining to filter B. Thus:

$$V_o = \frac{V_n V_f}{(R_c'/R_d')(R_d/R_c)V_d + [1 - (R_c'/R_d')(R_d/R_c)]V_f} \qquad \ldots\ldots (3)$$

Provided that:

$$(R_c/R_d) = (R_c'/R_d') \qquad \ldots\ldots (4)$$

equation (3) reduces to:

$$V_o = V_n V_f/V_d \qquad \ldots\ldots (5)$$

Assuming that equation (4) can be met, the validity of equation (5) depends upon the ability of the comparator and pulse generator to maintain the output of filter A constant at V_f. In practice this condition can be met to within any desired degree of accuracy by having sufficient loop gain and in the final circuit this constancy was maintained to better than 0·1 per cent.

* *Royal College of Advanced Technology, Salford.*

Variable Mark-to-Space Pulse Generator

The circuit of the variable mark-to-space generator is given in Fig. 3. Capacitor C is charged via a resistor R_1 and the resulting exponential voltage rise causes an exponentially rising current to flow through the tunnel diode TD_1 which is connected between the base and emitter of the germanium transistor VT_8. For tunnel diode currents below the peak point current the diode voltage is less than 50mV which is sufficient to maintain VT_8 switched off. When the capacitor voltage has risen to a value that the tunnel diode current reaches its peak point value the diode voltage rises rapidly to about 320mV thereby switching on VT_8 which discharges C via the small current limiting resistor R_2. By a suitable choice of component values the capacitor voltage will fall until the diode current is below the valley point when the diode voltage will fall rapidly to below 50mV thereby switching off VT_8 so that the cycle can recommence. The capacitor exponential voltage is coupled to the base of VT_9 in whose emitter circuit is another tunnel diode TD_2. The forward current through TD_2 equals the emitter current of VT_9 minus a control current i_c derived from the comparator. If the control current is constant and the emitter current of VT_9, i_e, is rising exponentially from zero, the condition for the voltage across TD_2 to rise and switch on VT_{10} is given by:

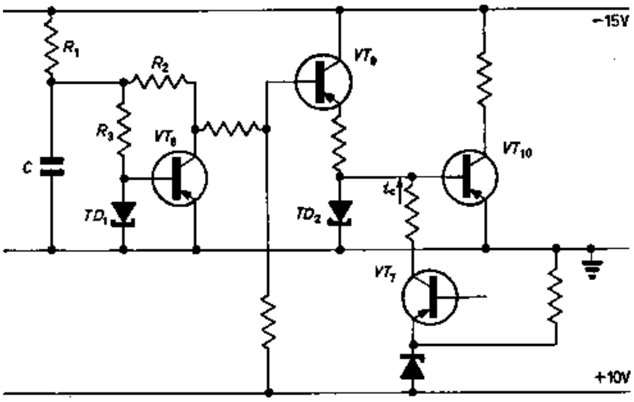


Fig. 3. Circuit of variable mark-to-space pulse generator

$$i_e = i_{pk} + i_c \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

where i_{pk} denotes the peak point current for TD_2. Since i_e is rising with the voltage across C the time at which TD_2 switches can be controlled by the magnitude of the control current i_c. When C discharges, i_e falls rapidly and eventually TD_2 switches back to its low voltage state, thereby switching off VT_{10}. Thus a positive going rectangular pulse with fast rise and fall times, controlled by the switching speeds of TD_2 and VT_{10}, will appear at the collector of VT_{10} and will last for a time which increases as the control current decreases. The relationship between the pulse duration and the control current will depend upon the waveform across C but since the duration is effectively controlled by the comparator ensuring that the output of filter A remains constant this relationship is unimportant.

Complete Circuit

The circuit diagram of the complete divider is given in Fig. 4. The denominator voltage V_d is chopped by two chopping transistors VT_1 and VT_2 before being applied to the comparator via the filter. Two chopping transistors are used to ensure that the clamped chopper output voltage is about 1mV even when the input voltage is of the order of 10V. The output of filter A is applied to the base of

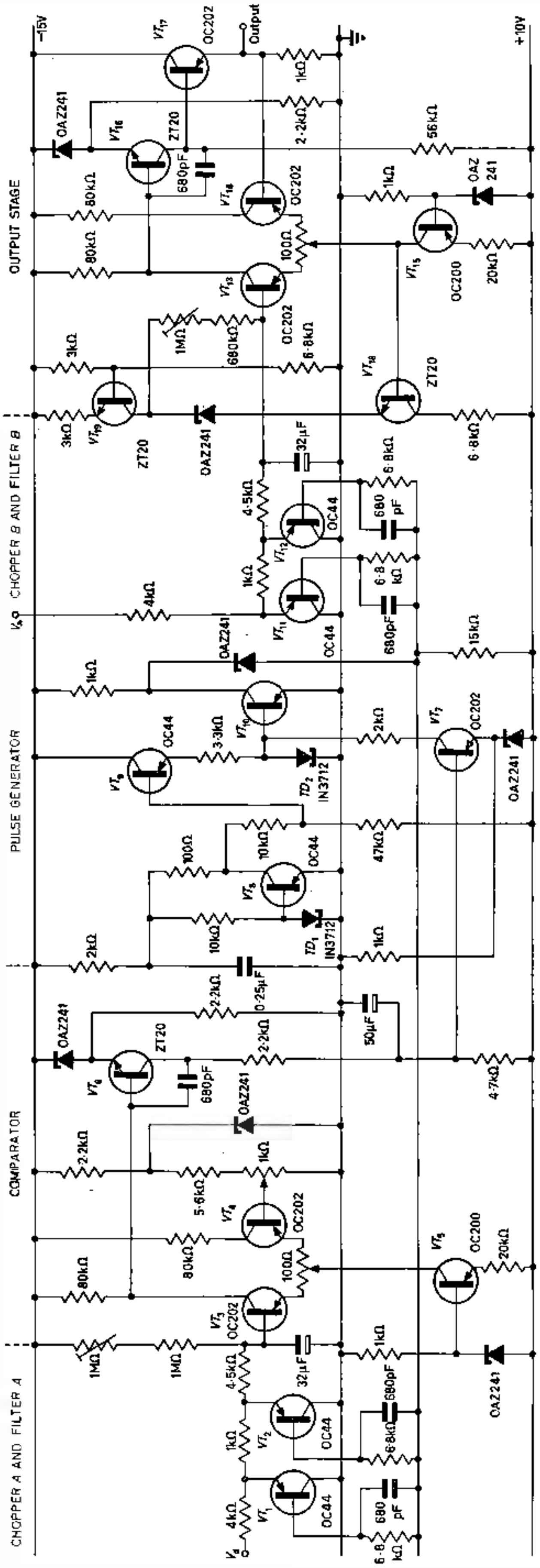


Fig. 4. Circuit of complete divider

VT_3 which with VT_4 forms a balanced long tailed pair amplifier employing a constant current generator (VT_5) in the common emitter circuit. The comparison voltage V_f, derived from a Zener stabilized supply, is applied to the base of VT_4. The output from the long tailed pair which is proportional to the error between V_f and the output of filter A is amplified further by VT_6 and VT_7 to appear as the control current for TD_2. A 2kΩ resistor is connected between the collector of VT_7 and TD_2 in order to limit the control current to a safe maximum value.

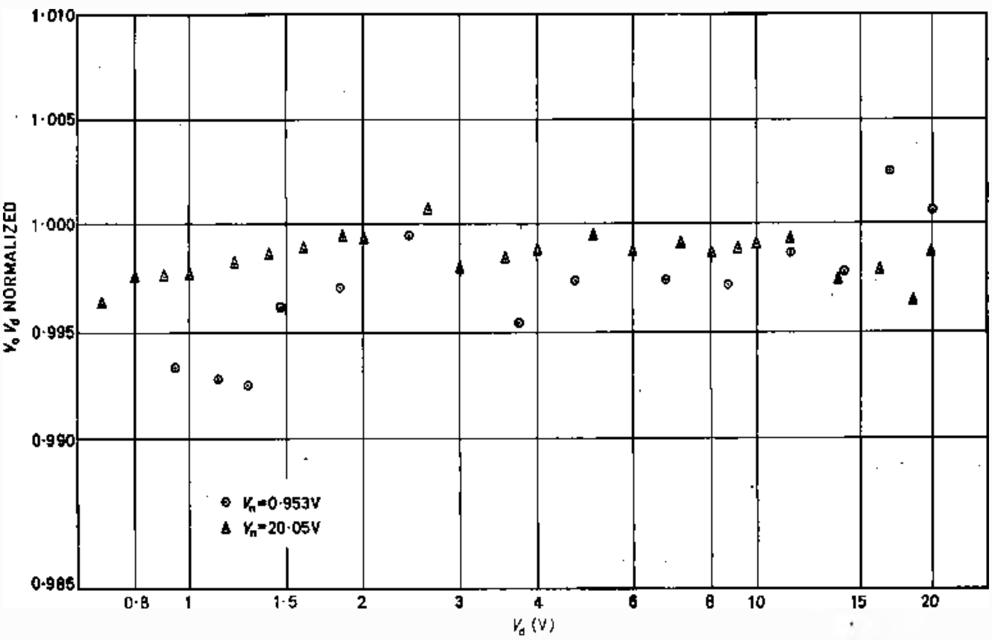


Fig. 5. Plot of normalized values of V_oV_d against V_d

The positive going variable duration pulse (10 to 600μsec at a p.r.f. of 1kc/s) at the collector of VT_{10} drives both sets of choppers, one for V_d and one for V_n. The divider output voltage appears across the capacitor of filter B.

This divider circuit will only work provided that no static current flows through the filter resistors. Such a current can arise from leakage currents from either the chopper transistors or filter capacitors: both, however, are negligible, at least at room temperatures. A further source of static current for filter A is provided by the base current of VT_3 which, however, can be diverted by a suitable resistor connected between the base and the negative line.

The divider output from filter B can be monitored on a high impedance voltmeter, but in the majority of applications this output will be required to drive another circuit. Therefore a further stage has been incorporated, of unity gain, high input impedance and low output impedance which acts as a sort of impedance transformer. The output of filter B is applied to the base of VT_{13} which with VT_{14} forms a long tailed pair amplifier. The output of this long tailed pair is amplified and applied to the base of VT_{14}. Provided that the long tailed pair is balanced the voltage at the base of VT_{14} exactly equals that at the base of VT_{13}. Because of the large amount of feedback the output impedance, taking the output from the base of VT_{14}, is very small, being of the order of a few ohms. In order to provide a high input impedance the base current of VT_{13} is diverted by a resistor to the Bootstrap circuit formed by VT_{18} and VT_{19}.

This impedance transformer also acts to increase the divider's accuracy at low output voltages. When the choppers are closed an offset voltage of about 1mV appears across them. Depending upon the pulse duration the filter outputs are in error by up to 1mV. If V_f is set at about 0·4V an error of $-0{\cdot}2$ per cent maximum is caused to the pulse duration. More important, however, is the error caused by the offset voltage to the divider output particularly when this output is small. Since a small output appears either when V_n is small or V_d large one can at least partially correct for the latter case by offsetting the output of the impedance transformer by 1mV. Under this regime the error due to chopper B offsets is given by $-0{\cdot}1/V_n$ per cent so that provided V_n is greater than 0·3V the overall error should not exceed 0·5 per cent.

A further source of error lies in the temperature dependence of the balance of the comparator and impedance transformer long tailed pair amplifiers. In an experimental divider the four unselected transistors were mounted in an aluminium block and the output voltage changed by $-0{\cdot}6$ per cent when the temperature was raised by 30°C giving a temperature coefficient of $-0{\cdot}02$ per cent/°C.

Performance

A sensitive test for the accuracy of a divider is to keep one input, V_n say, constant and vary V_d when the product V_oV_d should remain constant. Fig. 5 shows the results from two tests in which $V_n = 0{\cdot}953$V and 20·05V. Values of V_oV_d normalized to the expected value given by $V_n(V_f - 1\text{mV})$ have been plotted and in both cases the divider output never deviates by more than 0·8 per cent from the expected value—usually the deviation is nearer 0·4 per cent. Most of the experimental points lie below the expected value and this can be accounted for by the fact that the filter resistances were not exactly equal. An error of x per cent in the equivalence of R_o/R_d and R_o'/R_d' leads to an error in the output of $x(1 - V_f/V_d)$ per cent. Fig. 5 indicates that $x = 0{\cdot}2$ per cent and after allowing for this the divider error becomes of the order of $\pm 0{\cdot}2$ per cent which is of the same order of accuracy as that of the digital voltmeter used to measure the voltages.

The frequency response of the divider as built is poor but this could be improved, at some expense in accuracy, by reducing the filter time-constants which would introduce some ripple and also by increasing the p.r.f. above the 1kc/s used. This would reduce accuracy by making the chopped signals' rise times become more important.

The work described in this article forms part of a larger project supported by the Department of Scientific and Industrial Research.

Short News Items

The 1965 Physics Exhibition organized by the Institute of Physics and the Physical Society is to be held at the Manchester College of Science and Technology on 5 to 8 April.

This Exhibition, which is the 49th in the series originally held in 1905 by the Physical Society, is the first to be held outside London.

Admission is by tickets only, which are obtainable free of charge on application to the Secretary of the Institute and the Society at 47 Belgrave Square, London S.W.1.

The Handbook is now available price 10/- (post 2/- extra).

The 1966 Physics Exhibition will be held in London at Alexandra Palace from 28 to 31 March and the 1967 Exhibition will be held in the same place from 17 to 20 April. The 1968 or 1969 Physics Exhibition may again be held in the provinces.

St. Thomas's Hospital Medical School has recently taken delivery of the first complete computer of average transients manufactured in the U.K. by Intertechnique of Paris.

This system is being used for the enhancement of signal-to-noise ratios in electron spin resonance spectra. The complete equipment incorporates the new version of the SA40B 400 channel pulse height analyser as well as a crystal controlled timing system, magnetic tape spectrum stripper and integrator.

The system operates on the principle of accepting the d.c. signal from the spectrometer and scanning it n times maintaining constancy of analysis. As a result of this scanning mechanism the noise increases by $\sqrt{n}$ but the signal increases by n. Accordingly, one can achieve a $\sqrt{n}$ gain in signal-to-noise ratio, and hence be able to find signals which, in some cases, may originally be well below the average noise level.

There are many applications of this technique covering electron spin resonance, nuclear magnetic resonance, electrocardiology and virtually any other application where a constant signal is being sought in the midst of random noise.

The Compagnie Francaise De Television (CFT) and the Societa Electronica Italiana (SELIT) are jointly developing a new picture tube for colour television reception.

This new tube is claimed to give a brighter and better quality picture and can be used by all the existing colour television systems.

The Scientific Instrument Manufacturers' Association (S.I.M.A.) is to hold its fifth Group Exhibition in Moscow at the Polytechnic Museum on 26 May to 4 June, when 32 of the association's members will be exhibiting.

In addition to the exhibition there will be a lecture programme. Support for the Exhibition, organized by S.I.M.A. with the co-operation of the British Embassy in Moscow, is being given by the Board of Trade.

G.E.C. (Telecommunications) Ltd has been awarded a £350 000 contract for the provision of a microwave radio relay system in the Central American republic of El Salvador. The system will provide a capacity of 300 telephone channels in the 7Gc/s frequency band and will connect the towns of San Miguel, Santa Ana, Sonsonate and Usulatan with the capital city, San Salvador.

The equipment being supplied to the National Telecommunications Administration (ANTEL) is the latest completely semi-conductored type, no thermionic valves or tubes being used.

The equipment will be manufactured at the Company's plant in Coventry, England.

The 1966 British Joint Computer Conference will be held at the Congress Theatre, Eastbourne, Sussex on 3 to 4 May 1966. Sponsoring the conference, under the aegis of the United Kingdom Automation Council, are

The British Computer Society
The British Institute of Management
The Institute of Chartered Accountants
The Institution of Electrical Engineers
The Institution of Electronic and Radio Engineers.

The Organizing Committee, under the Chairmanship of Professor Stanley Gill, Imperial College, London, has provisionally agreed on the following topics for inclusion in the conference programme:

Scientific Applications
Management planning and decision making
Current problems
Engineering of real time systems
Future horizons

Contributions of up to 3 000 words are invited for inclusion in the programme. Those wishing to offer material are asked to send a brief synopsis of their proposed contribution to the Conference Secretariat, I.E.E., Savoy Place, London. W.C.2 by 31 July 1965.

Further details and registration forms will be available in due course from the I.E.E.

The Marconi Co Ltd has received an order valued at over £½M from the Sui Northern Gas Pipelines Ltd, of Karachi, for the engineering, supply and installation of the telecommunications, telemetry and instrumentation for the first phase of a new natural gas pipeline to be laid in the northern part of West Pakistan.

Civil engineering construction of the first phase, over 300 miles long, of this new pipeline is already under way. It will run from the present pipeline terminal at Multan, which is fed from one of the world's largest natural gas fields at Sui, 70 miles north of Sukkur, to the major centres of industry and population of North Pakistan such as Lyallpur and Lahore.

Closely following the route of the pipeline will be a comprehensive Marconi v.h.f. multi-channel communications system providing full telemetry, telephone and teleprinter facilities between all offices and stations on the pipeline. In the case of a complete trunk radio failure, a standby h.f. radio system will also link the main centres. When the installation is complete, the pipeline engineers will be able to instantly monitor, supervise and control the whole pipeline operation from the main central control room at Lyallpur.

At the moment the Marconi Company is carrying out a radio survey of the whole system and it is planned to have the complete telecontrol, data handling and telecommunications network fully operational in 19 months' time. Marconi engineers will undertake the training of Pakistani engineers on the equipment in the field, and also at the Marconi College and at the main Marconi Works at Chelmsford, England.

Madrid's Central Post Office is to be equipped with automatic mail handling systems that will start operating in October of this year.

Two sets of machines, each comprising letter culling and letter-facing units, have been ordered from Standard Elektrik Lorenz A.G. of Stuttgart, the German associate company of Standard Telephones and Cables Ltd.

The new machines will speed mail through Madrid's principal sorting office and are able to process letters and postcards and cancel stamps at a speed of 20 000 to 30 000 per hour.

Morganite Resistors Ltd of Jarrow has just obtained a new order from

America for 22 million resistors worth approximately $175 000.

These components, which will be delivered in twelve weeks, are for use in colour television, radio and other electronic fields, and in U.S. military equipment.

The Instrument Department of Ferranti Ltd at Moston, Manchester, has received a contract from Messrs. Brown-Boveri and Company of Mannheim, Germany, to supply the summation metering equipment for the Mangla Dam Power Station in West Pakistan.

Summation metering equipment is designed to measure the individual loads on the various feeders, with reactive power if needed, the total energy consumption of all feeders (subdivided if thought desirable) and a printed record of the half-hourly demand.

The main feature of the Ferranti equipment is the replacement of electro-mechanical summation by electronic methods. The conventional mechanical contactors fitted to watthour meters are replaced by photo-electric devices and the check and total registers are driven by an electromagnetic stepping motor, thereby obviating the need for mechanical links and reducing wear and maintenance to a minimum.

The summator—which is completely compatible with electromechanical equipment—can accommodate up to 18 display registers, i.e. 16 channels and 2 totalizers of variations on lower number of channels.

The entire equipment is transistorized, with all the circuit functions being provided on printed cards, so that the required cubicle space is reduced to less than a quarter of that occupied by conventional electro-mechanical equipment.

G.E.C. (Electronics) Ltd has supplied the Experimental Navigation Division of the Ministry of Aviation with high accuracy recording equipment for use by the Aeroplane and Armament Experimental Division Establishment at Boscombe Down in the assessment trials of aircraft navigational systems.

The data recording system is installed in the Division's newly acquired Comet 4C aircraft and is in regular use for inter-service testing and evaluation of all types of navigational systems and electronics.

The airborne digital data recorder contains a ferrite buffer store which is used to sample and hold 32, 18-bit words of data prior to punching out on paper tape. During a typical operation the recorder samples 27 digital inputs in 1·8msec and records the information on tape at 100 characters/sec. This is repeated at rates of up to 20 per minute throughout the flight. The data recorder is of modular construction, enabling additions and modifications to be made for different trials, as required. It uses a standardized digital interface compatible with other airborne digital equipment produced by G.E.C. Electronics.

The British Broadcasting Corporation has placed a contract with EMI Electronics Ltd to supply and erect a Band III aerial at Winter Hill, near Bolton, to improve BBC 1 television coverage in West Lancashire. The aerial will be co-sited with other aerials on the Independent Television Authority's new 1 000ft mast being erected at this site.

Erected at a mean mast height of 780ft on the triangular section of the mast, the 40ft aperture aerial will consist of four rings of full-wave dipole panels, each comprising two panels facing in north-easterly and north-westerly directions. It will be completely enclosed in a 12ft diameter cylinder of glass-fibre reinforced plastic for protection against weather and to facilitate all-the-year-round maintenance.

Two independent halves of the aerial —each of which continues to operate if a fault develops in the other—will each be fed by a 3⅛in diameter semi-flexible transmission line from the transmitters.

Vertically polarized, the aerial will transmit on Channel 12. The maximum effective radiated power is 100kW, directional over an arc of 180°, with reduced power in other directions.

Standard Telephon of Vienna, an associate company of Standard Telephones and Cables Ltd, of London, has supplied a new electronic signalling and telephone system for installation on the Dachstein Cableways in Austria.

The new system uses the car traction cables themselves to carry the precision signalling currents and telephone calls. These signals are applied via an induction transformer through which the cable passes without making any contact.

Maintenance and servicing costs have been substantially reduced as a result of introducing the new electronic system, which was operated very successfully for several months alongside the previous signalling system before being brought into full service.

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA) announce an increase in the exports of valves and semiconductors for the last quarter of 1964 based on Customs and Excise returns.

The increase in value, noticeable under all categories, counteracted the overall decline in the earlier quarters, bringing the total for the three months to approximately £3¾M, 23 per cent greater than in the first quarter, which was the previous highest.

This has brought the total exports for 1964 to nearly £11·4M a marginal increase over the 1963 figure of £11¼M.

The Annual Dinner of the Institution of Electrical Engineers was held at Grosvenor House on Thursday, 25 February 1965, when the President, Mr. O. W. Humphreys, C.B.E., was in the Chair.

The principal guest was Mr. R. J. F. Burrows, M.A., LL.B., President of the Law Society, who proposed the toast of 'The Institution', to which the President replied. The toast of 'Our Guests' was proposed by Sir Albert Mumford, K.B.E. (Immediate Past-President), and the reply was given by Professor C. B. Buchanan, C.B.E., Professor of Transport, Imperial College of Sciences and Technology.

Elliott Traffic Automation Ltd, a member of the Elliott-Automation Group, has set up a Transportation Survey Department to carry out transportation surveys and associated work for local authorities, major industrial organizations and public transport undertakings.

The new Department will work in close collaboration with Birmingham University's Departments of Transportation and Environmental Planning under Professor J. Kolbuszewski.

This Transportation Survey Department will help to meet the urgent needs of Borough and County Councils for up-to-date information on which to assess present and future transportation requirements in their areas. This problem has become increasingly important since the publication of the Buchanan Report and the Government's declared intention to establish a co-ordinated national transportation policy.

It will also be of considerable help to port authorities and other major industrial undertakings, which need to establish the effect of changing traffic patterns on current operating procedures, and to public transport undertakings which often require assistance with optimum scheduling and time-tabling procedures.

The Society of Non-Destructive Examination is to hold a Summer Meeting at Nottingham University on 8 to 10 July on "Automation in Non-destructive Testing." These will be two full sessions with discussion periods and four papers will be presented.

The Conference is open to non-members. Applications should be made as soon as possible to:

Hon. Secretary, Mr. G. A. K. Stobie, Corrosion & Welding Engineering Ltd, Kingston Road, Leatherhead, Surrey.

Correction. In the article 'LOPAD—A Logarithmic On-Line Data Processing System for Analogue Data' on pages 146 to 151 of the March issue, the following additions and corrections should be made.

To caption of Fig. 5, add *See Results. Fig. 6 caption should read '(c) z-output. To equation (9) add 'c.' before expressions contained in large parentheses.

LETTERS TO THE EDITOR

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

'D.C.' Amplifiers

Dear Sir,—I think it would generally be agreed that the term d.c. amplifier is now accepted as the abbreviation for 'directly coupled' amplifier. Because the abbreviation d.c. has been in use since long before directly coupled amplifiers were even thought of as the abbreviation for direct current, confusion is often caused and even worse, fundamental errors in design requirements can result in the use of this term. For example, a chopper type amplifier used for amplifying direct signals is not directly coupled but a.c. coupled and it is not therefore a d.c. amplifier even though it may be used for amplifying d.c. signals. (I am being deliberately facetious here).

It is high time that a new term was used to cover all amplifiers used to amplify direct signals whether they be currents or voltages, so what better could be used than direct signal amplifier which may be abbreviated to d.s. amplifier.

M. R. Harknett,

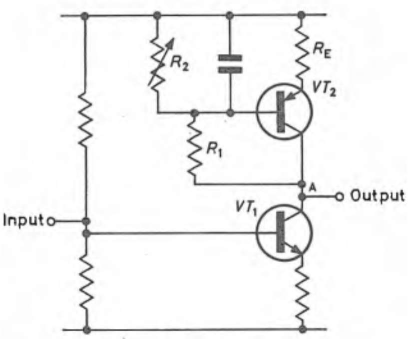
Department of Electronics,
The University, Southampton.

A Novel Integrator

Dear Sir,—A variant of the circuit described by Mr. Bertoya in your November issue, which I have used with remarkable success, might interest your readers.

It was required to design an audio transistor amplifier with an output stage having relatively large d.c. bias current and at the same time high output impedance to drive a mean rectifier circuit. The conventional method is to use an inductor but the circuit shown in Fig. 1 achieves the same purpose at much less cost.

Fig. 1. The integrator circuit



The a.c. collector load of the transistor VT_1 is

$$R_{ac} = R_1$$

The d.c. collector load, ignoring the V_{be} of transistor VT_2, is

$$R_{dc} = (R_1/\beta) + R_e + (R_1R_e/R_2)$$

β refers to the common emitter large signal current gain of transistor VT_2.

R_2 is normally very large and is varied to adjust the voltage at the point A. Hence, approximately,

$$R_{dc} = (R_1/\beta) + R_e$$

A d.c. load of 500Ω with an a.c. load of 20kΩ are readily obtainable with this circuit.

Yours faithfully,

S. Ghosh,

Standard Telephones & Cables Ltd.,
Basildon, Essex.

A Direct Coupled Negative Impedance Convertor

Dear Sir,—An active circuit of increasing importance in the realization of given transfer functions is the negative impedance convertor (n.i.c.). For instance, Linvill[1], Yanagisawa[2], Hudson[3], and others have described its application in deriving wave filter characteristics from *RC* networks. When a low-pass response is required, it is necessary for the n.i.c. to function down to zero frequency. The circuit therefore becomes subject to drift in working level and, in addition, a number of trimming adjustments may have to be incorporated to compensate for 'parasitic' impedances. Drew and Gorski-Popiel[4] give an example in which, around two transistors, there are a Zener diode, six fixed and five preset variable resistors.

A rather simpler direct coupled circuit may be derived from that due to Larky[5] by the introduction of an emitter coupled pair in place of the input transistor, as shown in Fig. 1. The voltages at A and B are equal and of like polarity, whereas the currents are of opposite sense. The circuit is stable when terminals B are short-circuited so, in use, the admittance, $1/Z$, presented to them would be arranged to exceed that presented to A.

R_1 is adjusted so that, when A and B are both short-circuited to earth, they carry zero current (i.e. only the standing base currents flow in $R_{2,6}$).

R_2 adjusts the impedance conversion ratio.

It can be shown that, if the base currents of VT_1 and VT_2 may be neglected, A presents an impedance very close to:

$$-\frac{G-1}{G+1}\cdot(R_2/R_6)\,(Z-(R_6/G-1))$$

where

Z = load connected to B

G = gain at C with respect to differential input A to B.

The necessary value of G can be obtained from the knowledge that an x per cent change in G gives rise to approximately a $2/G\ x$ per cent change in impedance conversion ratio. Resistor R_4 provides feedback to reduce variation in G and in the working level at *C*.

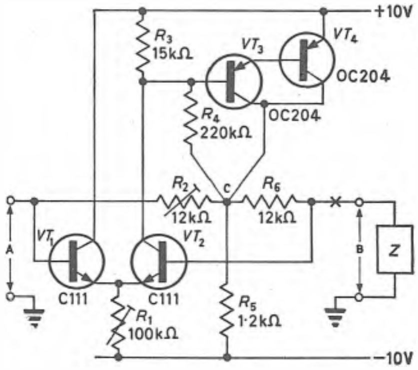


Fig. 1. The negative impedance convertor

Additional stabilization may be achieved by making R_3 and/or R_4 temperature sensitive.

The residual series term in the above expression can often be absorbed into Z but, if necessary, a resistor of value $R_6/G-1$ may be inserted between the n.i.c. and terminal B (X in Fig. 1). The resultant, although of reduced magnitude, now becomes rather sensitive to variations in G. Overcompensation also leads to A becoming short-circuit stable instead of B.

The voltage swing at C is equal to $|1+(R_6/Z)|$ times that at B, so it is advisable to make $R_{2,6}$ as low as practicable.

D. P. Franklin,

Electrical & Musical Industries Ltd.
Study and Assessment Group,
Wells Division.

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3. Hudson, —. *IBM J. Res. Developm.* (Jan. 1963).
4. Drew, —, Gorski-Popiel, —. Letter to Editor. *Proc. Instn. Elect. Engrs. 111* (1964).
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BOOK REVIEWS

Basic Electronic Circuits

Combined edition, 430 pp. Royal 8vo. Technical Press. 1964. Price 42s.

THE Services train large numbers of technicians for the routine maintenance of electronic equipment. The mathematical background of the trainees is normally very limited and teaching methods need to avoid the use of mathematics wherever possible. This leads to the presentation of facts in a pictorial form so they can readily be linked with practical devices. The book under review was developed by the Royal Electrical and Mechanical Engineers (R.E.M.E) in conjunction with the Technical Training Command of the R.A.F. and Decca Radar Ltd. The book is wholly concerned with pulse and waveform-shaping circuits, associated topics such as voltage stabilizers, digital and logic circuits are not included.

Part 1 of the book deals with the transient response of *CR*, *LR* and *LCR* circuits, the delay line and regenerative trigger-circuits. Circuit diagrams and waveforms are given showing the use of both valves and transistors. Unfortunately, some of the circuits are not particularly well designed, for instance on p. 95 a monostable multivibrator is shown with an OC75 transistor having a base resistor of 1MΩ to h.t. and a collector load of 10kΩ, i.e. the minimum β would need to be 100. There is no consistent policy concerning waveforms, in some cases these have been drawn in diagrammatic form, in others photographs from a cathode-ray tube are used. The diagrams mostly show the d.c. voltage levels and the relative phase between the various waveforms. However, the c.r.o. photographs do not show the levels and often the phase-relations are not indicated. A further difficulty is that in at least two cases the photographs have been inverted. In a manual of this type, accurate, complete and clear waveform diagrams are essential and the careless presentation of waveforms detracts from the usefulness of the book. Part 2 of the book covers valve and transistor circuits for timebase generation and strobe pulse markers. Here the circuits given seem to be properly designed and all waveforms are shown as line diagrams complete with d.c. levels and correct phase-relations.

The text of the book is in the form of short notes and is mainly concerned with supplementing the information in the circuit diagrams. Numerical calculation of such things as time-constant, scantime and current flow are not included.

Books attempting to cater for wholly non-mathematical readers are few and it is clear that a great deal of work has gone into the preparation of this manual. At this elementary level the presentation of facts in a pictorial form is an excellent way of teaching and an electronic serviceman or wireman could undoubtedly benefit from following the scheme of this book. However, it is neither suitable for (or intended to be) a vehicle for serious study with examinations in mind.

V. H. Attree

Semiconductor Network Analysis and Design

By Vasil Uzunoglu. 372 pp. Med. 8 vo. McGraw Hill Book Co. 1964. Price $12.50.

THIS book aims at providing on the one hand a thorough understanding of the operation of control circuits which use semiconductor devices, and on the other to give sufficient information on the design of such circuits required for their efficient operation under static and transient conditions.

The first four chapters bring a brief discussion of the physics of semiconductor devices and their intrinsic characteristics in terms of hybrid parameters. The remaining twelve chapters form the main part of the book, discussing in detail circuit applications of semiconductor devices, mainly transistors: low- and high-frequency, d.c. and operational amplifiers, oscillators, switching and negative resistance devices such as tunnel diodes, four-layer pnpn diodes, thyristors etc., with special attention paid to feedback and stability considerations.

Many of these topics are discussed in great detail, and free use is made of the special mathematical methods of network analysis. There is no doubt that the book will be read with profit by highly qualified electronic engineers or specialists experienced in circuit theory and control engineering. However, in the preface the author states that "this book can be considered also as an ideal text book on the graduate level. The reader will find that some basic knowledge of network theory, pole-zero concepts, Laplace transforms and determinants will aid in the reading of the book." This to the reviewer's mind is an understatement: unless the reader is fully conversant with all these topics—any many others—he will soon be left behind in the study of this book. Such items as Bode diagrams, poles and zeros, Butterworth- and Chebychev-functions etc. etc. are used very freely throughout, without any introduction apart from a very few explanatory lines (on p. 123, after first having been mentioned some fifty pages previously). Some symbols and characteristics are used without any explanation or definition; thus one is left to guess that $\mathcal{L}$ stands for Laplace transform; "immittance" is used without any definition etc. These shortcomings are probably due to the author's great familiarity with the subject, but they do not help the student who does not possess this highly specialized background training. It is therefore suggested that in any future edition some pages be added to introduce the well-trained though not specialized University graduate to some of the special analytical techniques used throughout the book. This might be done, if necessary, even at the cost of omitting the first four chapters as little use is made of them in the main part of the book.

With these reservations the book can be highly recommended.

E. Billig

Field Effect Transistor Applications

By William Gosling. 166 pp. Demy 8vo. Heywood Books. 1964. Price 40s.

THIS book is one of the first to deal solely with the field effect transistor or f.e.t. Although this transistor is not new, it is only recently that the planar manufacturing process has made the manufacture of these devices much easier and given a device with a superior performance to the alloy junction type. The planar process has also enabled a new transistor to be made, known as the m.o.s.t., which has a completely insulated gate. However, both these very important types get scant treatment in this book, especially the application using the enhancement mode of operation that is possible using a m.o.s.t. It is not made sufficiently clear that both p and n channel units are available and in chapter 5 a change is made in the circuits described to the complementary f.e.t. for no apparent reason.

To designers using 'ordinary' bipolar transistors the effect of temperature on the transistor characteristics is very important. Unfortunately the treatment of this topic is rather spread out within the book. The description of noise receives only a single page to itself and, while radiation damage is mentioned in the introduction, it is ignored in the rest of the book.

The major portion of the book is devoted to an analysis of linear amplifier circuits. This is clear and straightforward with no mysterious mathematical jumps. Considerable emphasis is put on using a combination of bipolar and field-effect transistors. This is a very valuable treatise in view of the growing emphasis on solid-state circuit design.

In the chapter on non-linear circuits a description is given of frequency convertors and demodulators. However, it is rather surprising that no specific description is given in the amplifier chapters on i.f. or audio amplifiers and the use of the high input impedance property in electrometer circuits. The chapter on non-linear circuits deals with many topics, but misses out the specific application of f.e.t.'s to micro-power circuits.

A most interesting and useful chapter is that on using the f.e.t. as a voltage controlled resistor. This is an application which has a large potential for the f.e.t. and it is particularly well done.

To those familiar with valve techniques some of the circuits and their treatment will be very familiar. This book will obviously be of value to those who have worked mostly with bipolar transistors and who now wish to include in their circuits field effect transistors. To these people, particularly those designing linear amplifiers, this should prove a useful introductory book.

R. E. Hayes

Cold Cathode Tube Circuit Design

By D. M. Neale. 259 pp. Demy 8vo. Chapman Hall Ltd. 1964. Price 45s.

IT is always surprising to find that so little comprehensive literature exists dealing with the applications of cold cathode tubes. The history of these devices is one of the longest of all the branches of electronics—gas discharges being one of the first phenomena studied and described. Many papers have been published dealing with the physics of gas discharges and some applications but this book is the first to attempt to present a comprehensive review of cold cathode tubes and circuits. The first two chapters trace the history of cold cathode tubes and deal generally with the physics of the gas discharge, subsequently all the more important groups of tubes, reference and stabilizer tubes, trigger tubes, counting tubes and display tubes are discussed in turn. Each section has a description of the class of tube and its construction and then considers the various applications. Numerous worked examples of design are given and this is a particularly valuable aspect of the book enabling circuit designers to achieve good designs with minimum of effort. Each chapter has a number of useful references which enable the reader to follow up, in greater detail, any particular point of interest and there is also a series of tables giving data and equivalents for a wide range of cold cathode tube devices. This book is ideal for use in any electronic laboratory and forms a valuable reference source book for the electronic engineer generally.

D. Reaney

Modern Control Theory

By J. T. Tou. 427 pp. Med. 8vo. McGraw-Hill Book Co. 1964. Price 112s. 6d.

THE linear, the non-linear, the adaptive, the noisy, the single and the multi-looped, the continuous and the incremental, all are given remarkably coherent treatment in this 'modern' book. Perhaps not for the old-school control engineers who content themselves with stodgily-applied network theory, but undoubtedly for the thrusting newcomers, this work provides a wealthy compendium of powerful computer-orientated theories of great promise.

After two introductory chapters of philosophical and mathematical indoctrination the reader is made ready to appreciate the discussion in chapter 3 on the concept and use of state-space as a method of analysis for continuous and discrete-data systems, leading to design methods in chapter 4. Chapter 5 shows the use of variational calculus in optimal control problems; and the principle of maximum, and its application to system design, follows in chapter 6. Dynamic programming is the subject of the next chapter, and lastly chapter 8 elucidates the theory of computer-control of processes.

Three factors are made quite apparent. Firstly, the important realization that the crucial problem in control system design is one of optimization. Secondly, that the design of such systems can be done so much better using computers than by struggling with blunt pencils, and thirdly that Professor Tou is a master of his art. Throughout the text good illustrative examples are used, and also there are many unworked ones left to test the reader's ability. Some 152 valuable references of recent and classic work relevant to the discussion are appended, providing the student with excellent sources of information.

The choice of the description, 'modern', in the title is acceptably apt in 1965, but adjectives such as 'advanced', 'difficult', or even just 'more' would have been equally apposite. This book should be read by all those who wish to remain well informed in advanced control engineering theory.

K. C. Garner

Transistor Bandpass Amplifiers

By W. Th. Hetterscheid. 314 pp. Philips Technical Library. 1964. Price 76s.

THIS is the first book to be devoted to a detailed study of the various aspects of transistor band-pass amplifier theory. It contains much original material that has not appeared in book form before. By treating the transistor as a two-port network as far as external behaviour is concerned, it is ensured that the results obtained can be readily extended to apply to any one of the common emitter, common base and common collector configurations. The use of y- and h-matrix parameters is particularly emphasized. It is, however, regrettable that these two sets of parameters are not related to the hybrid-π equivalent circuit.

The size of the book and its mathematical content are clear indications of the relative complexity of the subject. It gives detailed accounts of the following important aspects of transistor band-pass amplifiers: stability, tuning procedures, skew in the selectivity characteristic, neutralization, and cascading when the amplifier involves single tuned and double tuned networks. Also the important effects of transistor parameter spreads and coupling transformer imperfection on the stability performance of the amplifier are examined. Although a number of illustrative examples are included, the design and construction aspects of the subject are considered in a second volume that is to appear later.

The book seems to be free from gross errors, and it is a most useful addition to the available literature on transistor circuits. It would be particularly welcomed by the analytically minded electronics engineer concerned with transistor bandpass amplifiers.

S. S. Hakim

British Special Quality Valves and Electron Tube Devices Data Annual

By G. W. A. Dummer and J. Mackenzie Robertson. 1391 pp. Demy 4to. Pergamon Press. 1964. Price £10

This volume supersedes the British Miniature and Subminiature Valves Data Annual and provides designers and users with complete data on a wide range of special quality valves.

Included also is information on special electron tubes and devices such as cathode-ray tubes, photo-cells, counters, electro-luminescent panels, klystrons, magnetrons and stabilizers.

Guide Technique De L'Electronique Professionnelle 2 Volumes

Les Guides Techniques Industriels. 590 + 760 pp. Demy 4to. Publicite et Editions Techniques. 4th Edition. 1964/5. Price 160 F.

This guide, published in English, German and French is an informative source of reference to the French electronic industry.

In addition to the conventional lists of manufacturers and products, the guide gives details of the structure and growth of the French electronic industry, together with information on its latest achievements.

A list of importers of foreign equipment is also included.

American Microelectronics Data Annual 1964-65

By G. W. A. Dummer and J. Mackenzie Robertson. 941 pp. Demy 4to. Pergamon Press. 1964. Price £8

This Data Annual contains full information on thin film, semiconductor integrated and hybrid circuit assemblies at present available from American sources.

A special index enabling location of assemblies by circuit function or design category is provided in addition to the main index.

Masurari Electronice In Industrie

By T. H. Nicolav. 452 pp. Med. 8vo. Editura Technica Bucuresti. 1964.

This book written by seven authors is an introduction to the science of electronic measurement in laboratory and industrial applications, and is designed to assist the practical engineer faced with the problem of making various industrial measurements.

A Guide to Advanced Electrical Engineering

By R. V. Buckley. 167 pp. Crown 8vo. English Universities Press Ltd. 1964. Price 12s. 6d.

It has been the aim of the author, in preparing this book, to present to students studying the subject of advanced electrical engineering and in particular to those students preparing for the Institution of Electrical Engineers Part III Examinations a text-book of problems, both worked and unworked, together with a brief theoretical introduction to the subject.

Excitation Control

By G. M. Ulanov. 100 pp. Demy 8vo. Pergamon Press. 1964. Price 30s.

The author's aim is to present the reader with the principles of control associated with the use of excitation as a controlling or regulating effect. Automatic systems employing such an effect are some of the most important devices in the field of regulation and control, and, although this fact has been recognized, the theoretical development of these systems is lagging behind that of principles of deviation control. This book will be of interest to design engineers and industrial organizations working in the automatic control field. As a supplement to textbooks it will also be useful to control engineering students in their final year.

Worked Examples in Advanced Electrical Engineering

By K. S. Chapman. 158 pp. Demy 8vo. Edward Arnold Ltd. 1964. Price 21s.

This book was written to meet the needs of students preparing for Part III of the Examination of the Institution of Electrical Engineers in the subject of Advanced Electrical Engineering. It should also prove useful to those students studying for Engineering Degrees, Diploma in Technology, post-Higher National Diploma courses leading to a particular College award, and similar advanced courses in advanced electrical engineering.

Worked Examples in Alternating Current for Engineering Students

By P. J. Freeman. 108 pp. Demy 8vo. Macdonald & Co. 1964. Price 40s.

This book is divided into eight sections and contains eighty-four worked examples and thirty-four diagrams. It deals with basic single- and three-phase circuit theory, power factor correction, the single-phase transformer and the three-phase induction motor and is intended for use by students in the following courses: 1st Year Engineering Degree; 1st Year Diploma in Technology (Engineering); 1st Year Higher National Diploma; 02 National Certificate; T3 City and Guilds Electrical Technicians.

Vocabolario Di Televisione

By A. Sabbatini. 190 pp. Med. 8vo. Stampato dalla Litografia Mandolini. 1964. Price L.2 000

This is a dictionary of television terms and expressions in Italian, English, French and German, and contains in addition a cross-reference four language index.

Optoelectronic Devices and Circuits

By S. Weber. 360 pp. Demy 4to. McGraw-Hill Book Co. 1964. Price £6

This book is a compilation of 97 significant articles from 'Electronics' covering various aspects of optoelectronics.

In twelve chapters, this volume covers: modern optoelectronic principles; laser design and application; optical communications; military and space applications; infra-red systems; display systems; pattern recognition; computers and digital applications; industrial applications; instrumentation; new concepts in television; and unconventional optoelectronic devices.

Industrial Electronics Measurement

By A. Schure. 128 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 15s.

The book, originally published in the U.S.A., has been so written that it can be fully understood by engineers, as well as by those in whose province the instruments truly lie. It is largely descriptive, and mathematics have been kept to the essential minimum.

RCA Receiving Tube Manual

By RCA. 608 pp. Med. 8vo. Radio Corporation of America. 1964. Price $1.25

This Manual, like its preceding editions, has been prepared to assist those who work or experiment with home-entertainment-type valves and circuits.

The material in this edition has been augmented and revised to include the recent technological advances in the electronics field. Many valve types widely used in the design of new electronic equipment only a few years ago are now chiefly of interest for renewal purposes. Consequently, in the Tube Types Section, information on many older types is limited to basic essential data; information on newer and more important types is given in greater detail.

Modern Digital Circuits

By S. Weber. 358 pp. Crown 4to. McGraw-Hill Book Co. 1964. Price 74s.

This book contains 109 articles which have been published in 'Electronics' during the years 1961 to 1963 and dealing with digital circuit design.

The Digital Computer

By K. C. Parton. 125 pp. Crown 8vo. Pergamon Press. 1964. Price 17s. 6d.

The object of this book is to describe as concisely as possible how to use a modern computer, the sort of work already being carried out to advantage, how the computer itself actually works and, finally, the considerations necessary to ensure a ready acceptance of computer help in an existing organization.

An Introductory Course for Electrical Technicians

By R. J. Hartles. 222 pp. Demy 8vo. Sir Isaac Pitman. 1964. Price 18s.

This book deals with the electrical principles which are taught in introductory courses such as the First and Second Years of the Electrical Technicians' Course of the City and Guilds of London Institute.

This book includes sections on graph plotting, practical work reports, and the writing of technical descriptions, and concludes with a summary of the quantities which have been employed, with their units, abbreviations and defining relationships.

ELECTRONIC EQUIPMENT

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

(Voir page 275 pour la traduction en français; Deutsche Übersetzung Seite 282)

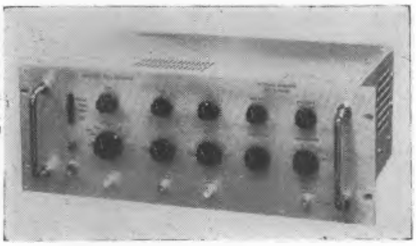
DOUBLE PULSE GENERATOR

Distributed by: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Illustrated below)

E-H Research Laboratories of Oakland, California, are currently manufacturing their model 132L pulse generator in England. Positive or negative, single or double phase pulses of 50V peak amplitude into a load of 50Ω are generated, with rise and fall times of 10nsec.

The model 132L may be operated in either triggered or free running modes from 10c/s to 3·5Mc/s as a single pulse



generator or 10c/s to 1Mc/s as a double pulse generator. A modification may be installed at extra cost to permit operation down to 1c/s.

Pulse width and delay are continuously variable from 100nsec to 10msec by front panel controls.

Complex pulse patterns may be generated by using the internal gate circuits—the application of a −5V d.c. signal to a front panel socket inhibits the output pulses for the duration of its presence.

The timing circuit of the 132L is entirely transistorized. A significant feature of the circuits employed in the design of the instrument is the freedom from all pre-set waveform adjustments—output pulses are extremely clean and fast with no interaction between amplitude, frequency, width or delay controls.

EE 80 751 *for further details*

AUTOMATIC DEWPOINT INDICATOR

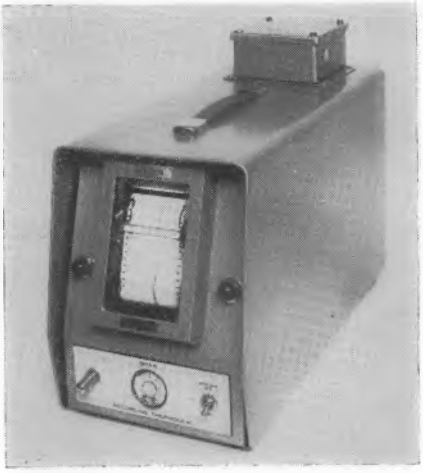
Shaw Moisture Meters, Rawson Road, Westgate, Bradford, 7, Yorkshire

(Illustrated on right)

Known as the 'Thermodew' this instrument gives absolute readings of the dewpoint temperature of air or gas using the new thermocooled semiconductors, latest type photocell, and thermistor temperature measurement, it provides an automatic, self-contained, thermocooled dewpoint meter. A recording version is also available.

The accuracy of the dewpoint temperature reading is inherent in the design. The light from a lamp, operated from a stabilized voltage supply is reflected on to a photocell, when dew is formed on a mirror surface. The current to the thermo-module used to cool the mirror then ceases, so that the mirror surface, of which the temperature is measured, continually cycles between the dew and no dew conditions. This prevents drift in any of the circuits from causing cooling of the mirror below the dewpoint temperatures. An indicator light on the front panel confirms that the dew—no dew, condition is being maintained. This method of operation is considered a significant technical advance over the servo systems which have been used previously in the automatic measurement of dewpoint temperature. Solid state components are used throughout.

The standard model has a time clock incorporated which has an adjustable 'on' time, and an adjustable interval between periods of operation. This is normally one minute in each ten minute period. Should a dewpoint reading be required before the clock is due to switch on the cooling circuit, a biased toggle switch (marked Auto.off.on) is pressed down manually, until a steady reading of dewpoint is indicated on the dial. The reason for this method is to give the 'Thermodew' a reliable performance and to ensure that it is not left switched on continuously when not required. Several years trouble free operation can be expected.



A vibrator type air pump is fitted at the rear so that the dewpoint of ambient air can be indicated. The pump, which is also controlled by the time switch, purges the measurement chamber once in 10min. The pump sucks in air through the sintered bronze filter at the rear. If the Thermodew is required to check gas piped to it, then the flexible tube is removed from the inlet connector. There are inlet and outlet gas connectors to the test chamber and these can be coupled to a gas supply which should have a slight pressure to maintain a small flow, as the pump is no longer in circuit.

EE 80 752 *for further details*

INSTRUMENTATION TAPE RECORDER

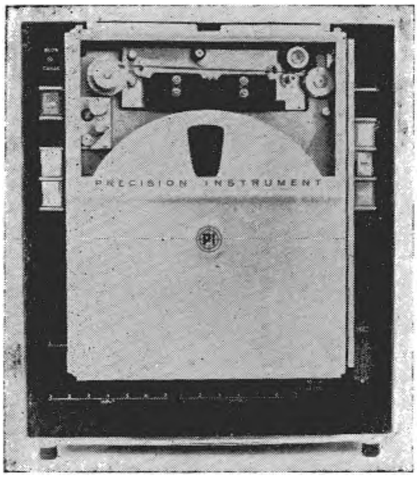
Distributed by: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustrated on page 269)

The main feature of the Precision Instrument Co. Ltd model PI-2100 instrumentation tape recorder is its ability to provide an operating performance to standards normally only associated with much larger and more expensive laboratory equiment. Its performance is in fact, of sufficiently high standards to enable it to be used for 99 per cent of likely data recording assignments, with the added advantage of such small size and portability that 'on site' operation in many otherwise impossible environments is quite feasible. This convenience is further supplemented by facilities for remote control up to a distance of 100ft, and for operation from a variety of power sources including batteries and 28V d.c. supplies as well as a.c. sources of various frequencies. Another feature which adds to the convenience of operation is a specially designed tape-loading magazine which enables a change of tape to be made in a few seconds, including the automatic threading of the tape over the heads. There is also provision for continuous loop magazines up to 24ft.

Two versions are available, a 7 channel model and a 14 channel model, and each model is available with either 4 speeds or 6 speeds. The speed range on the 4 speed versions is 60, 30, 15 and 7½in/sec, and the range on the 6 speed versions is 60, 30, 15, 7½, 3¾ and 1⅞in/sec. A photoelectric device automatically provides shut-off at the end of the tape run, but all functions can then be made operable by means of a special 'overide' push-button control.

Switching speeds by means of the front panel push button controls in both the recording and the playback modes automatically switches in the appropriate electronic circuit which is all solid state. Thus the need for costly time consuming re-calibration or re-adjustment is entirely eliminated.



Facilities for either direct or f.m. recording or reproducing are available on all channels, the frequency response with direct operation being extended to 300kc/s at 60in/sec and with f.m. operation to 20kc/s.

Flutter is less than 0·6 per cent peak-to-peak over a bandwidth of 0·2c/s to 20kc/s at 60in/sec and flutter compensation is available as an optional extra.

In order to ensure the highest possible mechanical stability, the head assemblies, tape guides, and capstan are all mounted on a single plate of stress-relieved aluminium 1¼in thick, and the ground and lapped surface of this plate, flat to a tolerance of 5 tenths of a thousandth of an inch, mates with the precision ground mounting surfaces of the head-assemblies. All track widths, spacings, numberings and tolerances conform with IRIG standards.

The PI-2100 will operate at temperatures from 40°F to 120°F and in conditions up to 95 per cent relative humidity, non condensing. It is provided with shock and vibration mountings which will meet all military specifications.

EE 80 753 *for further details*

ROTARY SAMPLING SWITCH

I.D.M. Electronics Ltd, 1 Arkwright Road, Reading, Berkshire

This high speed rotary sampling switch has been designed by I.D.M. Electronics Ltd to fulfil many of the requirements demanded by remotely operated systems. It can be used to solve many of the problems usually associated with telemetery work, as d.c. signal chopping, waveform generation, amplifier stabilization, low level sampling of strain gauge and thermocouple signals, digital counters, and in other high speed or low speed applications where information is sampled.

The combination of noble metal contact materials, wiper materials and insulation materials have been developed to produce the optimum performance of very low electrical noise, low thermal e.m.f. low driving torque, high insulation and long life. The mechanical design incorporates preloaded precision ball races ensuring consistent performance over the whole life of the switch. The construction of the switch permits it to be one of the smallest of its kind, and particular care has been taken to extend the life of the switch without maintenance, and without the use of oil-baths, etc.

Pulse positional accuracy is of the highest order enabling units to be used with automatic de-multiplexing equipment. In high frequency systems where inter-coupling between channels is difficult to eradicate this type of switch provides the minimum possible coupling between segments.

The design permits a wide range of configurations such that the product of the number of poles times the number of ways equals 24 (e.g. 1 pole 24 way, 2 pole 12 way, etc.). The switch can be driven by a miniature a.c. or d.c. motor and as the switch bank requires only a few gramme-centimetres torque even at high speed the power consumption is very low.

Segments may be interconnected to provide a large variety of mark-to-space ratios, or if so desired, special track configurations will be considered. Lead-wires can be supplied plain, colour-coded, or can be screened and in either case they are left for attachment to the customers terminal block.

EE 80 754 *for further details*

SOUND LEVEL METER

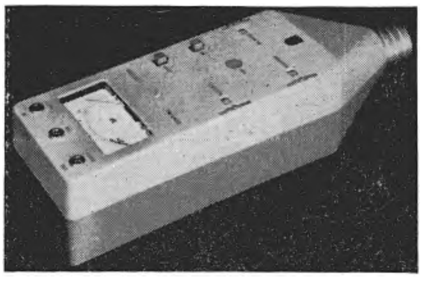
The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(Illustrated below)

The Philips sound level meter type PM6400 has been introduced by The M.E.L. Equipment Co. Ltd as a basic instrument for use by engineers, architects, medical workers and others concerned with acoustic measurement and the control of noise.

It is completely transistorized and in consequence is very compact and weighs less than 6 lb. This portability and the fact that operation is from internal batteries greatly facilitate measurements both in the laboratory and on site.

An omnidirectional moving-coil microphone is fitted at one end of the instrument, and sound level is indicated on a moving-coil meter calibrated from −7



to +10dB. Sockets are provided for the connexion of a recorder or a filter unit.

The instrument conforms to the I.E.C. Standards, and covers the range 33 to 130dB. The accuracy at 1kc/s, and with the instrument at right-angles to the source of sound, is plus or minus 1dB. The frequency range is 25c/s to 15kc/s and I.E.C. standard A, B and C networks are incorporated.

Power is supplied by two nickel cadmium cells which provide 20 hours operation before requiring charging. A battery charger type PM9000 is also available.

Dimensions are 13½ × 4½ × 3¼in. An 'ever ready' leather carrying case is supplied with the instrument.

EE 80 755 *for further details*

R.F. SIGNAL GENERATOR

K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(Illustrated below)

The G.30 is a very convenient and inexpensive signal generator, covering the frequencies from 120kc/s to 240Mc/s in seven switched bands. This wide coverage makes it useful for most experimental and alignment work. The r.f. output which is in excess of 100mV can be internally or externally modulated, the internal modulation being supplied at 400c/s. The depth of modulation is adjustable from 0 to 50 per cent by a knob control. The 400c/s modulating frequency is also available at two sockets on the front panel.

EE 80 756 *for further details*

THYRISTOR TRIGGER CONTROL UNITS

Standard Telephones & Cables Ltd, Rectifier Equipment Division, Edinburgh Way, Harlow, Essex

(Illustrated on page 270)

Single- and three-phase thyristor trigger control units, now available from STC, provide a variable delay firing pulse to the gate of thyristor(s) for such applications as power supply regulation, d.c. motor speed control and many other circuits requiring voltage or current control. The units are generously rated and are designed for long life and trouble free operation.

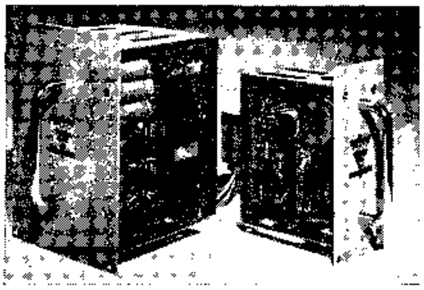
STC thyristor trigger control units are versatile electronic assemblies supplied as compact units for panel mounting flush into the fascias of silicon rectifier equipment. An alternative assembly allows for the chassis mounting of the trigger units inside the cubicles. Connexions are via plugs and sockets or via terminal blocks if preferred.

The principal features of these units are:

(1) Self-contained power supply—only a.c. mains input required.

(2) Fast time response.

(3) Square wave firing signal gives full control for inductive circuits.
(4) Pulse height unaffected by mains voltage changes of +20 per cent and −50 per cent.



(5) Will control thyristors down to 2 per cent of maximum output voltage.
(6) Can be used with all standard rectifier connexions.

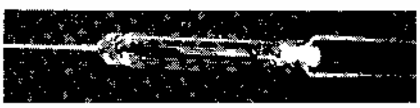
EE 80 757 *for further details*

DRY REED SWITCHES

B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustrated below)

The latest Gordos dry reed switch available from B & R Relays Ltd, is a mechanically biased changeover switch. Known as type MR201 it has an overall length of 3·09in, glass length 1·2in and a glass diameter of 0·2in. The contact rating is 10VA at a maximum of 0·25A. A new contact material is used on the type MR201 called 'Di-met'. 'Di-met', which is a proprietary name, gives plated contacts of dissimilar metals. It combines gold for stability, copper for low resistance and added alloys for hardness and durability. These prevent 'sticking' or 'cold welds'. Features of 'Di-met' are the extremely low contact resistance which it provides, and good 'inrush' capacity for switched tungsten lamps and similar loads. It is more inherently stable than such interim contact materials as rhodium and silver. Plating thickness is kept to a minimum, but allows for



control of drop-out or 'operating differential'. In appearance one blade is silver in colour and the other copper-gold.

The full range of Gordos dry reed switches can now be supplied with either 'Di-met' or diffused gold contacts.

In addition to other uses the new MR201 can be incorporated in B & R's R03 and R04 dry reed relays.

EE 80 758 *for further details*

MINIATURE TRANSFORMERS

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

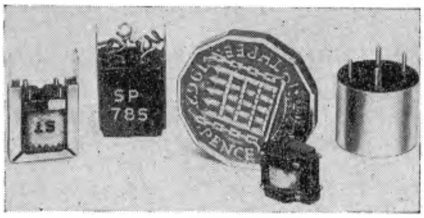
(Illustrated above right)

A comprehensive range of miniature transformers and chokes suitable for pulse, audio, power and instrument transformers and close tolerance inductors and chokes is available from Parmeko Ltd.

This range, known as the 4 000 series, has been awarded certificates of Advanced Quality Approval in the Humidity Class of H2 and Temperature Category of 40/70 (RCS 241) by the Joint Service Radio Components Standardization Committee. General construction of these transformers and chokes is in accordance with DEF 5214.

Open models from the 4 000 series utilize the complete Parmeko range of single and double bobbins and are particularly suitable for applications where compact reliability is required. Most of these open models are available with ferrite cores. Terminations are made to pins moulded into the bobbin cheeks. Three types of clamp can be supplied to suit individual mounting requirements.

Three alternative forms of assembly are available: encapsulated, aluminium cased, and magnetically screened. The encapsulated models are resin cast and have the additional advantages of greater mechanical rigidity and electrical insulation. The aluminium cased potted models employ epoxy resin or other suitable compounds to present a simple form of mounting with all the advantages of the



encapsulated models. The problems of magnetic interference are overcome by using Mumetal cases, and double or triple screened models can be supplied when required. The terminations of these alternative forms may be supplied to suitable pins or flying leads, depending on the form of assembly and the number of winding ends.

EE 80 759 *for further details*

PRINTED CIRCUIT MOTOR AMPLIFIERS

Pye Printed Motors Ltd, St. Andrews Road, Cambridge

(Illustrated on right)

The printed circuit motor, by reason of its design, poses problems in the design of suitable servo amplifiers. The motor is essentially a high current, low voltage device and therefore, if the full benefits of its design qualities are to be realized, the servo amplifier must be designed to suit.

An amplifier chain is now offered consisting of a power amplifier with its associated driver and operational amplifier, suitable for controlling the series 4 motor under all normal conditions, and the larger motors with restriction of the maximum speed (because of their higher back e.m.f.).

A short specification of a typical power amplifier P.A.1G and its associated driver amplifier D.A.1G is:

Power supply:
30V d.c. centre-tapped (±15V d.c.).
Continuous current:
8·75A max.
Voltage gain:
D.A.1G + P.A.1G − 2 times.
Nominal output impedance:
0·05Ω.
Max. dissipation at 25°C ambient:
34W.

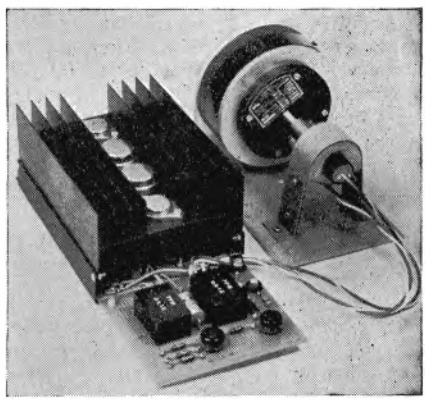
For applications requiring currents greater than 8·75A two or more power amplifier stages may be paralleled, and up to three of these can be driven by a single amplifier D.A.1G.

The operational amplifier O.A.1G is normally used with feedback, selected to define the phase gain characteristics of the servo loop. The amplifier has the following typical specification:

Power supply:
24V centre-tapped (±12V d.c.).
Quiescent current drain:
4·5mA.
Nominal open loop gain:
60 to 80dB.
Stable with 100 per cent feedback:
200c/s.
Open loop bandwidth:
200c/s.
Input voltage drift at source impedance:
1kΩ 500V/°C at 25°C.

The drive and operational amplifiers are modular blocks approximately 1¼ × 1 × ¾in high, suitable for mounting on printed circuit boards. The connector pins conform to a 0·1in matrix spacing.

The power amplifier is designed around an extruded finned heat sink. Its approximate overall dimensions are 6 × 4·25 × 3in.



EE 80 760 *for further details*

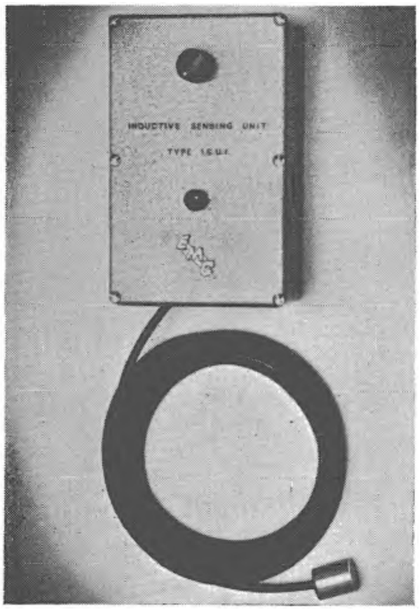
INDUCTIVE SENSING UNITS

Electronic Machine Control Ltd, Sherman Road, Bromley North, Kent

(Illustrated on page 271)

Electronic Machine Control Ltd is supplementing its existing ranges of photo-electric, ultrasonic and capacitance-type sensors with a new series of inductive units for industrial electronic control and counting systems. Available in a wide range of standard types, the new devices are non-contacting sensors influenced by the nearby presence of metal objects, both ferrous and non-ferrous. They are of robust construction, fully sealed and can be operated under water or in oils and non-corrosoive fluids.

Because of their simplicity—they have no moving parts—these sensors will function for long periods under severe conditions of temperature and vibration



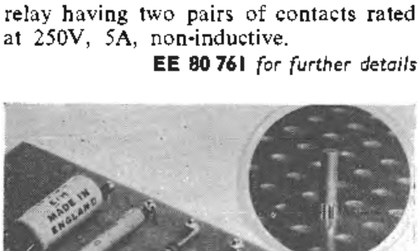
and require no maintenance attention in service.

The dimensions of a typical inductive sensor are as little as $1\frac{3}{4}$in by $\frac{7}{8}$in in diameter, and this compactness allows installation to be accomplished in confined spaces where other sensing methods would present difficulties or unnecessary complication. Any metallic mass, small or large, can be detected, even through glass or plastic materials.

A mains-operated electronic unit, measuring $4\frac{3}{4}$in × $7\frac{1}{2}$in × 2in, is supplied with each sensor and can be remotely mounted from it by as much as 50ft or greater. It is fully transistorized in the interests of long-term reliability, requires no warm-up time and, like the sensor itself, needs no maintenance. A single sensitivity control and indicator light are fitted on the front cover for initial setting-up purposes; once set, no further adjustments are required.

The electronic unit controls an internal relay having two pairs of contacts rated at 250V, 5A, non-inductive.

EE 80 761 *for further details*



TERMINAL PINS

Vero Electronics Ltd, South Mill Road, Southampton, Hampshire

(Illustrated above)

Vero Electronics Ltd has introduced a new terminal pin, part No. 2141, which can be used with any Veroboard (with or without copper strips) that has a hole diameter of 0·052in, or any printed wiring or terminal board drilled, or pierced with holes of the same diameter. The head and self-cutting serrations under it assure an extremely tight fit without fear of the pin falling out of the board during or after soldering. These new pins are manufactured from brass, with a tin-dipped finish to facilitate soldering.

EE 80 762 *for further details*

PLUG-IN COMPONENTS BOARDS

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustrated below)

Keyswitch Relays Ltd can now offer an inexpensive plug-in component board unit, the type 304, which enables circuit design engineers to use a common module for relays, miniaturized and solid state circuits, small motors and power packs.

The new unit comprises the 22-way moulded plug-in base of the Keyswitch P33 relay, with a circuit component board enclosed in a transparent dust-proof cover.

During production tests, as a demonstration of what can be achieved, Keyswitch successfully built into a standard unit a ring counter stage complete with power pack and comprising 11 trigger tubes, 11 diodes, 65 resistors and 35 capacitors, all within a mounting space of 3·125in × 2·218in × 1·093in. The Type 304 can also be supplied without the component board if desired.

Important features are ease of maintenance and modification, as units damaged by component or circuit failure can be replaced immediately from stock spares. Extra sockets can be mounted in existing equipment during initial manufacture, enabling subsequent modification to be made easily and economically.

The panel socket will accept only those units for which they are intended because of a non-interchangeable keying device, while spring-steel retaining clips in a selection of colours can be provided for coding and identification purposes.

Rating of the plug and socket assembly (mated) is 10A, with contact resistance 5mΩ and working voltage 750V a.c./d.c. maximum. Insulation resistance is better than 1 000Ω at 500V d.c.

The component board is of synthetic resin bonded paper laminate, to which are bonded longitudinal copper strips 0·0015in thick, spaced at 0·15in or 0·2in intervals. It is pierced with a regular matrix of holes. Wiring area is 2·875in × 2·218in, and current ratings (copper strip) range from 1A with temperature rise of 1·0°C to 5A with temperature rise of 39·0°C. Resistance is approximately 0·045Ω/in and insulation is better than 100MΩ. Maximum working voltage is 500V d.c. between adjacent copper strip.

EE 80 763 *for further details*

EDGE CONNECTORS

Ferranti Ltd, Kings Cross Road, Dundee, Scotland

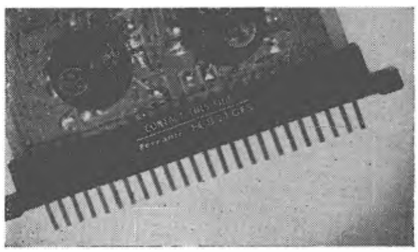
(Illustrated below)

In common with the growing range of electrical connectors produced by Ferranti Ltd printed circuit edge connectors feature a high degree of built-in reliability which is achieved by the use of a very low rate contact spring and of non-porous gold-plated contact faces.

Edge connector contact clips employ independent low rate springs which enable the contact forces to be accurately controlled to within close limits during manufacture. Clip and spring are of phosphor bronze, and the clip has a minimum of ·0002in of hard gold in the contact area. The mechanical design ensures that contact is made with the circuit board under all conditions with all specified thicknesses of board and with the correct force. Contact clips are formed from rolled wire strip to obviate sharp edges which could damage the board conductors, and as an additional safeguard, the contact faces are rounded.

As contact is made approximately 0·220in before the board is fully inserted, there is considerable latitude to inadequate board insertion and to misalignment.

Great care is taken to ensure that the contacts are gold plated to a very high and consistent standard. Samples from every batch are submitted to the DEF 5325 plating proving test, and they are also submitted to bending and burnishing tests. In addition, sections are taken so that the thickness of the plating in the contact area can be measured.



Statistical quality control methods are employed at all important stages of manufacture.

The connector mouldings are available either in glass-filled, flame-resisting, diallyl phthalate or high-grade, nylon-filled, phenolic material, the former material offering high dimensional stability at elevated temperatures. Sharp radii are avoided at all points where the board comes into contact with the moulding, and there is a chamfered lead-in to facilitate entry of the board. Polarizing keys are provided which may be inserted into any of the pole positions without removal of, or damage to, the contact.

These connectors are available in 8, 16, 24, 32 and 40 contact sizes with single- or double-sided contacts all with single solder spills. Contact spacing is 0·150in. The current carrying capacity is 1A per pole (at 100°C) and the d.c. working voltage is 450V. When used with an unchamfered printed circuit board, the insertion force is approximately 5·5oz per contact, but if a chamfered board is used the insertion force is approximately 3·5oz; the total insertion force is directly proportional to the number of poles.

The controlled contact pressure is sufficient to break down tarnish films, thus ensuring low contact resistance—less than 3mΩ—even at the very lowest working voltages.

EE 80 764 *for further details*

TEMPERATURE GAUGE

Startronic Ltd, 117a Malden Road, New Malden, Surrey

(Illustrated below)

This is a compact, portable battery-operated instrument for measurement of temperature from −20°C to +120°C in two ranges, −20°C to +55°C and +50°C to +120°C with an accuracy of ±1°C.

It uses fully interchangeable thermistors for temperature sensing.

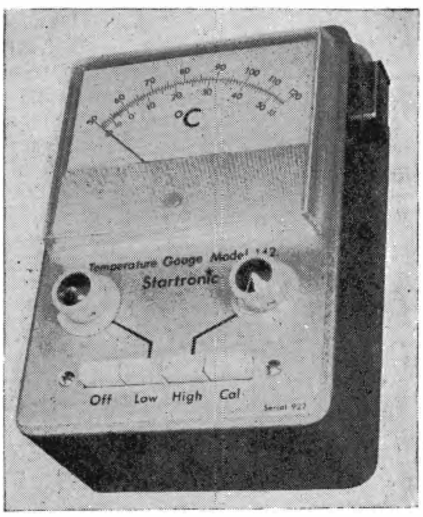
The scales have excellent linearity and the large meter provides a scale length of 3·75in.

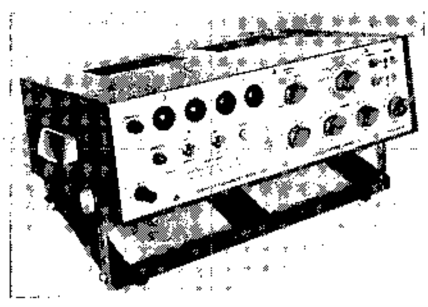
The instrument can be located a considerable distance away from the probe without seriously affecting the accuracy. For instance, a lead of 2Ω resistance gives an error of only ¼°C at 100°C.

Specially matched thermistors are available to enable the user to fix them where required in equipment for measurement of surface temperatures, and probes can be supplied for the measurement of gas or liquid temperatures.

In addition, a ten-way selector box and various extension leads are available.

EE 80 765 *for further details*





COUNTER-TIMER

J.A.C. Electronics Ltd, Station Estate, Blackwater, Camberley, Surrey

(Illustrated above)

J.A.C. Electronics Ltd is now in production with a new transistorized digital frequency and time interval meter which has been especially designed with a fully variable, crystal controlled time-base to provide a direct read-out in any predetermined units (e.g. rev/min). Its features include a high input impedance and an easy to read in-line display with an overspill indicator giving an effective seven figure read-out for four figure display cost.

Counter periods between 1msec and 10sec in millisecond steps are available for the measurement of frequencies from 0·1c/s to 300kc/s; the counting intervals may also be controlled manually or externally by signals and/or contact closures for the general determination of the number of events per unit time.

Magnetic or photo-electric (with integral lamp unit) probes are available for use in tachometry applications.

EE 80 766 *for further details*

PROCESS TIMERS

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated above right)

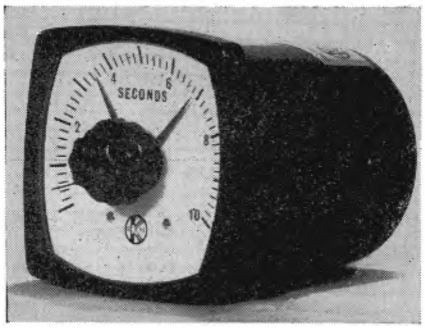
Kent Precision Electronics Ltd has produced a new range of solid state process timers for industrial application, as part of their range of solid state control instruments. The instruments are suitable for accurate electronic power timing in the range of 10msec up to 15sec with an overall accuracy of ±1 per cent.

Single or dual instruments which operate standard electromagnetic relays are housed in the approved standard panel-mounting instrument case as shown in the photograph. Instruments are also available with pulsed outputs to fire thyristor circuits for control of power up to the maximum available thyristor ratings.

The instruments consist basically of a stable *RC* timing circuit and an accurate transistor voltage threshold detector which switches when the voltage in the timing circuit reaches a definite value, and fires the appropriate thyristor output. For relay operation a low power thyristor is used to control a standard relay. For solid state control the output pulse is used to fire the appropriate thyristor stack.

One of the most attractive features of this range is the availability of a dual instrument within the standard panel-mounting case. This dual feature enables the instrument to be interconnected, so that after the first elapsed time occurs the second elapsed time is initiated, and the whole process may be recycled with settable on/off ratios. Alternatively, the dual instrument may be used as two separate timers with concentric controls.

EE 80 767 *for further details*



FREQUENCY SELECTIVE AMPLIFIER

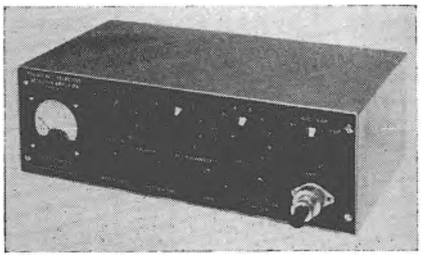
H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Illustrated on page 273)

H. Tinsley & Co. Ltd has recently introduced a new portable transistorized frequency selective amplifier. Battery operated, the instrument has many applications. It can, for example, be used advantageously with bridges or inductive voltage dividers, where measurements are required at a number of frequencies. Portability is facilitated by compactness and low weight. It has dimensions of 12in × 4in × 8½in and a weight of 5 lb.

The instrument is capable of detecting input signals of 0·1μV to 10V, and has a tunable frequency in four ranges covering from 10c/s to 100kc/s. The frequency is selected upon three decade dials adjustable to an accuracy of 1 per cent of frequency up to 10kc/s; with a *Q* factor greater than 30 and a sensitivity such that 0·1μV can be detected over the entre range.

Three stages of amplification are incorporated: using low noise, high gain transistors. The voltage gain of the amplifier is in the order of 10^6. This is not temperature dependent; due to the use of low leakage silicon transistors. The input impedance of the instrument varies with the degree of attenuation. The attenuator is divided into 11 steps of 10 to 120dB. For input signals greater



than 10μV, the input impedance is 10kΩ, with 10 to 50dB of attenuation; rising to 1MΩ, with 50 to 120dB of attenuation, for larger inputs.

An alternative model, type 5711, calibrated in angular frequency, is also available.

EE 80 768 *for further details*

PORTABLE THICKNESS GAUGE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

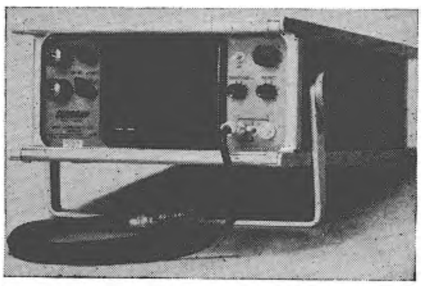
(Illustrated below)

The latest ultrasonic gauging equipment to become available from Dawe Instruments Ltd is the type 1805A Sonoray 30. It was designed primarily for battery operation in the field, yet has accuracy and resolution equal to those of most large production-line gauging units.

Ultrasonic gauging has been limited in the past by the need for mains power, which is not necessarily available, and cannot in any case conveniently be supplied from a trailing lead when the operator has to negotiate ladders, scaffolding, manholes and similar obstructions. The type 1805A is designed to overcome this problem. It is a compact (15 × 9½ × 4½in high), lightweight (16 lb complete), portable instrument normally powered by its own rechargeable battery. It can, however, be operated from the mains and is suitable for both flaw and thickness gauging.

The generator produces pulses at 1·5 to 5·0Mc/s, the reflections of which are recorded on a 2½ × 2½in screen. The scale can be expanded until the full width represents ½in thickness of steel. Gauging accuracy is ±0·005in for thicknesses in the range 0·1 to 6·0in. For flaw detection, where the need for accuracy is less critical, the range extends up to 100in.

Transducers must be matched to the frequency in use and are normally of the high-performance piezoelectric Z-type. The standard instrument is supplied with a 6ft transducer coupling cable, but cables of up to 200ft length may be employed without need for modifications. Straight beam, angle beam, through transmission and immersion testing may be used.



EE 80 769 *for further details*

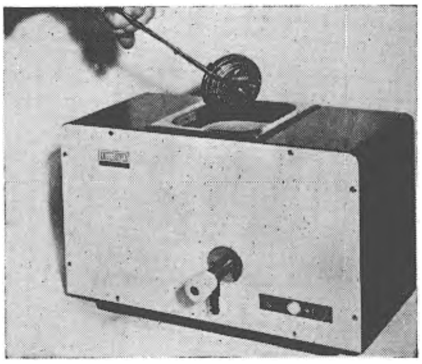
ULTRASONIC CLEANER

Sonics Division, Elliott-Automation Ltd, Bath Road, Beenham, Reading, Berkshire

(Illustrated below)

A new ultrasonic cleaner announced by the Sonics Division of Elliott-Automation is particularly suitable for the cleaning of delicate components in laboratories and precision industries and in the jewellery and watchmaking trades.

The equipment is extremely reliable and is completely self-contained. Both the 2-litre capacity tank (which measures 6in × 4½in × 5in internally) and the 60W generator are housed in a strong case measuring 14in × 8in × 9½in. A push-button on/off switch is the only control necessary as tuning has been eliminated by the employment of transistorized circuits in conjunction with the Elliott-Acoustica 'Multipower' transducer. The maximum cleaning efficiency is always available irrespective of the contents of the tank or changes in temperature. A



drain cock is provided to facilitate emptying the tank. A hose connexion on the drain enables a recirculating pump and filter unit to be used if required.

The equipment can be used by unskilled labour and the long-term stability of the power output enables cleaning times to be of standard duration.

EE 80 770 *for further details*

ULTRASONIC FLAW DETECTOR

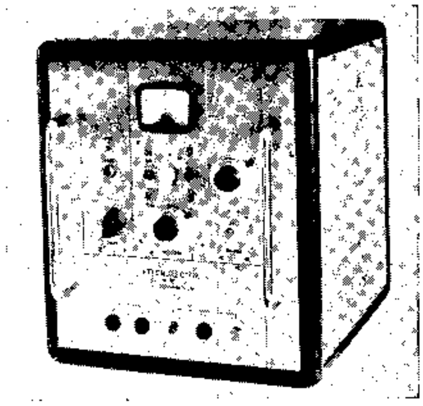
C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Illustrated above right)

This redesigned ultrasonic flaw detector, the Attenutector Mk. IV, is now available. It is designed for the non-destructive testing of plastics and graphite but is suitable for a variety of other applications such as the detection of bubbles in liquids, the checking of the bonding of plastic to plastic or plastic to metal, or metal to metal, and the variation of density in liquids.

The instrument, as its name implies, operates on a through transmission principle, where the ratio of attenuation through the test piece to the received signal can be as much as 1 000 000 to 1, consequently immersion techniques are prescribed to avoid any variation in contact between probes and specimen.

Standard probes are available for frequencies from 250 to 650kc/s and other



frequencies up to 2Mc/s can be provided to special order.

The system is in general more sensitive than X-ray inspection and is very much cheaper both in initial cost and running expense.

EE 80 771 *for further details*

ANALOGUE DIGITAL CONVERTOR

Moore, Reed & Co. Ltd, Woodman Works, Durnsford Road, London, S.W.19

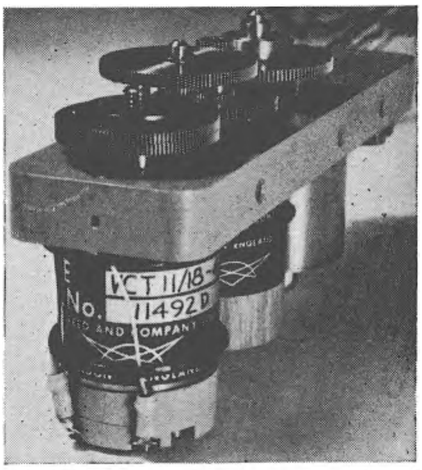
(Illustrated below)

A new addition to the range of control devices produced by Moore, Reed and Company Ltd, is an electromechanical convertor in which one step of a stepper-motor drive produces one digit change of an encoder output.

Applications include conversion of two-phase or three-phase a.c. signals into digital codes, binary or Gray code. Used with a subtractor, the unit will produce an analogue synchro, resolver or potentiometer output from a digital input. The convertor will respond to a pulse train from a programme unit, fed through transistorized logic to the stepper-motor driving the encoder, to give simultaneous analogue and digital signals, for control or indication purposes.

The use of precision gears incorporating anti-backlash techniques permits inaccuracies (stepper motor input/synchro output) to be reduced to the order of 5 minutes of arc.

The unit whose outside dimensions are approximately 5in × 1½in × 2¾in is available with binary V-scan, Gray or Gilham code digital outputs.



EE 80 772 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Date: 1 April
Lecture: Ranger 8 Spacecraft, with special reference to Moon Shot Camera Systems
By: B. P. Miller
Date: 29 April
Lecture: The Particles of Modern Physics (The 56th Kelvin Lecture)
By: R. O. Frisch

Electronics Division

Date: 2 April
Lecture: Features of a New Doppler VOR Beacon
By: C. W. Earp
Date: 5 April
Colloquium on: The Design of Real Time Computer Systems (Joint meeting with the I.E.R.E. Computer Group at 2.30 p.m.) (All wishing to attend must apply to the Secretary for tickets)
Date: 6 April
Lecture: The Future Impact of Integrated Circuitry on the Work of Professional Engineers (Joint meeting with the Science and General Division)
By: J. T. Kendall
Date: 7 April
Lecture: Assessing Precision Connector Pairs
By: I. A. Harris
Date: 9 April
Lecture: Electromyography (Joint meeting with the I.E.R.E. Medical and Biological Group at 2.30 p.m.)
Date: 12 April
Lecture: Solid Circuits in Computers (Joint meeting with the I.E.R.E. Computer Group) (Further details and particulars to be announced)
Date: 13 April Time: 2.30 and 5 p.m.
Colloquium on: Problems and Developments in Electron Linear Accelerators and associated Radio-Frequency Valves (Joint meeting with the Science and General Division) (All wishing to attend must apply to the Secretary for tickets)
Date: 14 April
Lecture: Acoustics and Telephone Transmission-Examples of the Problems of Human Judgment
By: H. S. Leman
Date: 26 April
Lecture: Standard and Convertors
By: P. Rainger, E. Rout
Date: 30 April
Lecture: Equipment and the Results of Tests with the NIMBUS Meteorological Satellite
By: N. E. Rider

Science and General Division

Date: 6 April
Lecture: The Future Impact of Integrated Circuitry on the Work of Professional Engineers (Joint meeting with the Electronics Division)
Date: 13 April Time: 2.30 and 5 p.m.
Colloquium on: Problems and Developments in Electron Linear Accelerators and Associated Radio Frequency Valves (Joint meetings with the Electrons Division) (All wishing to attend must apply to the Secretary for tickets)
Date: 21 April
Discussion: Have Inductively Coupled Ratio Bridges Superseded all other Bridges?
Opened by: R. Calvert
Date: 27 April
Lecture: A Travelling Field of Induction-Type Instruments and Motors
By: H. E. M. Barlow

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Computer Groups

Date: 5 April Time: 2.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Colloquium on: Design of Real Time Computer Systems
(Registration Necessary) (Joint I.E.E.-I.E.R.E.)

Radar Group

Date: 7 April Time: 6 p.m.
Held at: The I.E.R.E. Lecture Room, 9 Bedford Square, London W.C.1
Papers on: Inertial Navigational Systems for Airborne and Shipborne Uses. (Tickets will be required)

Television Group

Date: 14 April Time: 6 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Paper on: B.R.E.M.A. Colour Television Home Viewing Tests
By: R. N. Jackson, K. E. Johnson and B. J. Roberts
Date: 21 April Time: 6 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1
Paper on: Effect on the Ionosphere of Nuclear Explosions
By: F. L. Hill

Electro-Acoustics Group

Date: 28 April Time 6 p.m.
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Paper on: Synchronously Tuned Methods of Harmonic and Intermodulation Distortion Analysis (Tickets will be required)
By: D. E. O'N. Waddington

Medical Electronics Group

Date: 9 April Time: 2.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London W.C.2
Discussion: Electromyography (Joint I.E.E. I.E.R.E.)

Computer Group

Date: 12 April Time: 5.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2
Paper on: Solid Circuits in Computers (Joint I.E.E.-I.E.R.E.)
By: N. Miller

West Midland Section

Date: 6 April Time: 10 a.m.
Held at: University of Birmingham
Symposium on: Electronics in Industry—the next five years
(Joint meeting with the South Midland Centre Electronics Section) (Registration Necessary)

South Western Section

Date: 24 April Time: 7 p.m.
Held at: University of Bristol, Engineering Laboratories
Paper on: Microstructure and Physical Properties of Thin Film
(Joint meeting with the Institute of Physics and Physical Society)
By: A. J. Forty

South Midland Section

Date: 30 April Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham
Paper on: Integrated Circuits
By: G. C. Padwick

South Midland Section

Date: 1 April Time: 7 p.m.
Held at: Abbey Hotel, Abbey Road, Malvern
Paper on: Small High Fidelity Loudspeaker Systems
By: K. F. Russell

Yorkshire Section

Date: 14 April Time 6.30 p.m.
Held at: University of Leeds
Paper on: 405/625 Line Conversion Systems
By: C. R. Longman

Merseyside Section

Date: 14 April Time: 7.30 p.m.
Held at: Walker Art Gallery, Liverpool
Paper on: Field Effect Transistors and their Applications
By: C. S. den Brinker

North Eastern Section

Date: 14 April Time: 6 p.m.
Held at: Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle
Paper on: Pulse Modulation Systems
By: J. Balmer

THE TELEVISION SOCIETY

Meetings are held at the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 2 April
Lecture: Advanced Television Technical Problems in Japan
(Recorded Lecture)
By: K. Suzuki
Date: 29 April
Held at: The Royal Institute, 21 Albemarle Street, London W.1.
Lecture: The Specification of an Adequate Television Signal (The Fleming Memorial Lecture) (Tickets will be required)
By: R. D. A. Maurice

PUBLICATIONS RECEIVED

MULLARD INDUSTRIAL COMPONENTS is a new catalogue, the object of which is to provide engineers and designers with a quick guide to the Mullard range of components. The contents, indexed under their generic titles for easy reference, include the latest introductions covering modern design needs for industrial applications. Descriptions of the products are purposely abbreviated to provide the reader with a rapid appreciation of the salient characteristics, but reply paid cards are included to enable further details to easily be obtained. Requests for copies should be made on company headed paper to Central Enquiry Handling, Mullard Ltd, Mullard House, Torrington Place, London, W.C.1.

R.E.C.M.F. DIRECTORY—Produced by the R.E.C.M.F., this Directory is a comprehensive list of the Federation's Members and the products which they manufacture. An initial printing of 10,000 copies has been effected and arrangements are now in hand for the directory to be given world-wide distribution to overseas British Commercial Representatives, foreign Trade Associations, foreign Chambers of Commerce and leading overseas manufacturers. Further details of this directory are available from the Radio & Electronic Component Manufacturers Federation, 6, Hanover Street, London, W.1.

THE REPORT OF THE TIN RESEARCH INSTITUTE for 1964 is now available and gives details of various researches that may develop into useful outlets for this metal. Included in this report are details of the Institute's continued efforts to improve soldering processes by studying the surface treatments preceding soldering, especially in the case of copper foil which is the material that is soldered in printed circuits. This work shows gold, rhodium, and palladium coatings, applied by electrodeposition, to be ineffective and often deleterious whereas tinning the copper ensures superlative solderability. This report is published by the Tin Research Institute, Fraser Road, Perivale, Greenford, Middlesex.

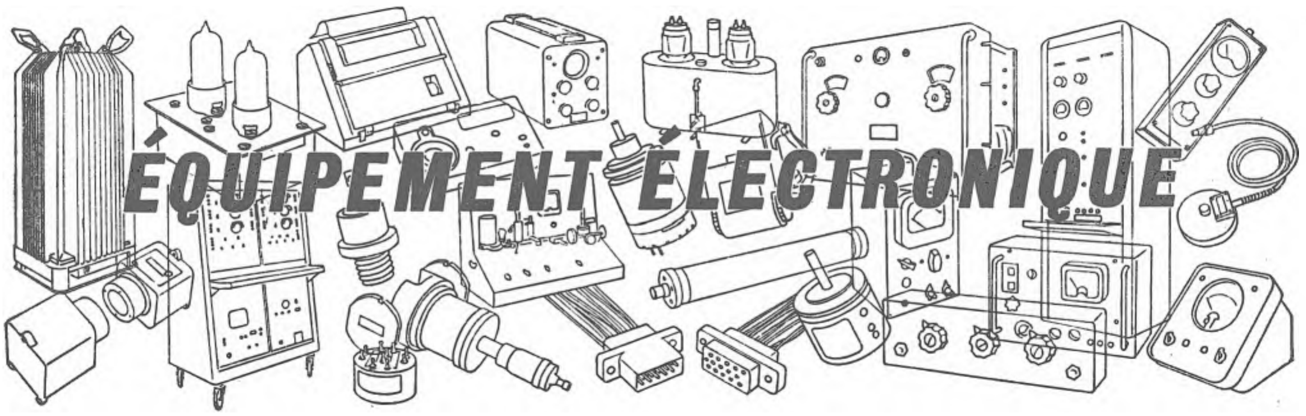
MEDIUM POWER HIGH SPEED SILICON DIODES are described in leaflets PD36-301 to PD36-305. These diodes have fast recovery times and cover currents from 6A to 35A. Copies of these leaflets are available from Westinghouse Brake and Signals Co. Ltd, 82, York Way, Kings' Cross, London, N.1.

THERMISTORS FOR SOLID STATE THERMAL SWITCHING is the title of a new STC leaflet (MK/189) which describes the PTC range of positive temperature co-efficient thermistors.

The new thermistor range is primarily for detecting excessive temperature in industrial equipment. The thermistors are very small spherical components (4·73mm diameter) that can be incorporated into electrical windings. In this way they are being used for thermal overload protection of the windings of motors, generators, transformers and electromagnets.

Performance details are given in leaflet MK/189 available from STC Semiconductor Division, Footscray, Sidcup, Kent. Telephone FOOtscray 3333. Telex 21836.

SELECTION CHART-PRECISION RESISTORS is a new booklet (issue 3) which provides detailed informaation in such a way that a resistor to meet a particular requirement can be quickly selected; the criterion may be one of resistance tolerance, dimensions, operation over a wide temperature range, excellent long term stability etc. ALMA loosewound resistors, the J Range and the Metal Film resistors are included in the tabulated lists of this latest edition. Requests for copies of this chart should be made to Alma Components Ltd, Park Road, Diss, Norfolk.



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 268 à 273

GÉNÉRATEUR D'IMPULSIONS DOUBLES

Distributeurs: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Illustration à la page 268)

La société E-H Research Laboratories d'Oakland, Californie, construit actuellement son générateur d'impulsions modèle 132L en Angleterre. Ce générateur produit des impulsions doubles ou simples, positives ou négatives, d'une amplitude de pointe de 50 V dans une charge de 50 Ω, avec temps de montée et de chute de 10 nsec.

Le modèle 132L peut être mis en action soit par le mode à déclenchement soit par le mode à course libre de 10 Hz à 3,5 MHz comme générateur d'impulsions simples ou de 10 Hz à 1 MHz comme générateur d'impulsions doubles. Il peut être modifié moyennant un supplément de prix pour permettre le fonctionnement jusqu'à 1 Hz.

La largeur et le retard d'impulsions sont à variation continue de 100 nsec à 10 nsec au moyen de commandes fixées sur le panneau frontal.

Des impulsions complexes peuvent être produites à l'aide des circuits de porte intérieurs: l'application d'un signal de tension continue de -5 V à une douille de panneau frontal refoule les impulsions de sortie pendant la durée de sa présence.

Le circuit de minutage du modèle 132L est entièrement transistorisé. Les circuits utilisés dans cet instrument se caractérisent par le fait qu'il rendent superflus tout ajustement préréglé de la forme d'onde: les impulsions de sortie sont extrêmement nettes et rapides sans interaction entre les commandes d'amplitude, de fréquence, de largeur ou de retard.

EE 80 751 *pour plus amples renseignements*

INDICATEUR AUTOMATIQUE DU POINT DE ROSÉE

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, 7, Yorkshire

(Illustration à la page 268)

L'indicateur "Thermodew" fournit des lectures absolues de la température au point de rosée de l'air ou du gaz. Par l'emploi des nouveaux semi-conducteurs à refroidissement thermique et d'une cellule photoélectrique du modèle le plus récent ainsi que de la mesure de la température par thermistor, il constitue un instrument de mesure du point de rosée automatique, autonome et à refroidissement thermique. Il existe également un version à enregistrement.

La précision de l'indication de température au point de rosée est inhérente à la réalisation de l'appareil. La lumière, émise par une lampe actionnée par une alimentation en tension stabilisée est réfléchie sur une cellule photoélectrique, provoquant ainsi la formation de rosée sur une surface de miroir. Le courant transmis au module thermique utilisé pour refroidir le miroir s'arrête à ce moment-là, de sorte que la surface du miroir dont on mesure la température effectue un cycle continu entre l'état de rosée et de non-rosée. Ce mouvement empêche la dérive dans n'importe lequel des circuits de provoquer le refroidissement du miroir au-dessous des températures au point de rosée. Un voyant lumineux sur le panneau frontal confirme le maintien de l'état de rosée ou de non-rosée. Cette méthode de fonctionnement est considérée comme un progrés technique important par rapport aux systèmes asservis utilisés jusqu'ici pour la mesure automatique de la température au point de rosée. L'indicateur est entièrement muni de composants constitués de corps solides.

Le modèle standard comprend une minuterie à temps de marche réglable et à intervalles réglables entre les périodes de fonctionnement. Cet intervalle est normalement d'une minute pour chaque période de 10 minutes. Si l'on veut obtenir une lecture du point de rosée avant que la minuterie n'effectue le branchement du circuit de refroidissement, on presse à la main un tumbler polarisé (marqué: "Auto.off.on) jusqu'à ce qu'une indication stable du point de rosée soit donnée sur le cadran. Le but de cette méthode est d'assurer au "Thermodew" un comportement sûr et d'empêcher qu'il ne reste en marche de manière continue lorsque cela n'est pas nécessaire. On peut s'attendre de la part de l'appareil à plusieurs années de fonctionnement sans la moindre panne.

Une pompe à air du type à vibrateur est fixée à l'arrière de manière à pouvoir indiquer le point de rosée de l'air ambiant. Cette pompe qui est également commandée par le commutateur de temps purge la chambre de mesure toutes les 10 minutes. La pompe aspire l'air à travers le filtre en bronze fritté à l'arrière. Si l'on veut que le Thermodew vérifie le gaz qui peut lui être insufflé, il y a lieu d'enlever le tube souple du connecteur d'entrée. Les connecteurs d'entrée et de sortie de gaz reliés à la chambre d'essai peuvent être accouplés à une alimentation en gaz qui doit être soumise à une légère pression pour maintenir un flux léger, la pompe n'étant plus en circuit.

EE 80 752 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera stand 61 Allée 61 au SALON INTERNATIONAL DES COMPOSANTS ÉLECTRONIQUES qui se tiendra à Paris du 8 au 13 avril 1965. Les visiteurs seront les bienvenus.

MAGNÉTOPHONE D'INSTRUMENTATION

Distributeurs: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustration à la page 269)

Le magnétophone d'instrumentation modèle PI-2100 de la société Precision Instrument Co. Ltd se distingue par une capacité de comportement à un degré normalement exigé d'appareils de laboratoire beaucoup plus grands et plus coûteux. Son comportement est, en fait, suffisamment élevée pour permettre de l'utiliser pour 99 % des travaux d'enregistrement de données. Il offre, en outre, l'avantage de dimensions extrêmement réduites et une "portabilité" qui permet de l'utiliser sur place dans des lieux où le fonctionnement serait autrement impossible. A cet atout s'ajoute la possibilité de télécommande jusqu'à une distance de 30,48 m et de fonctionnement à partir d'une variété de sources de puissance comprenant les batteries et l'alimentation en tension continue de 28 V ainsi qu'à partir de sources de courant alternatif de diverses fréquences. La commodité du fonctionnement est encore accrue par un chargeur de bande magnétique d'une conception spéciale, qui permet d'effectuer le changement de bande en quelques secondes, y compris la mise en place de la bande sur les têtes. On peut en outre fixer des chargeurs à bande continue mesurant jusqu'à 7,5 m.

Il existe deux versions du magnétophone: un modèle à 7 voies et un modèle à 14 voies, tous deux étant prévus pour 4 vitesses ou 6 vitesses. La gamme de vitesses sur les versions à 4 vitesses est de 152,4, 76,2, 38,1, et 19 cm/sec et celle sur la gamme des versions à 6 vitesses est de 152,4 76,2 38,1, 19, 9,52 et 4,76 cm/sec. Un dispositif photoélectrique assure l'arrêt automatique à l'extrémité de la bande mais toutes les fonctions peuvent alors être mises en action au moyen d'une commande à bouton-poussoir spécial.

La commutation des vitesses au moyen des commandes à bouton-poussoir du panneau frontal, tant pour l'enregistrement que pour le réenregistrement, branche automatiquement le circuit électronique approprié qui est entièrement constitué de corps solides. On élimine ainsi entièrement la nécessité d'un nouvel étalonnage occasionnant toujours une perte de temps coûteuse ou le réajustement.

Toutes les voies permettent l'enregistrement ou la reproduction directe ou en modulation de fréquence, la réponse de fréquence avec fonctionnement direct s'étendant à 300 kHz à 152,4 cm/sec et, sur fonctionnement à modulation de fréquence, jusqu'à 20 kHz.

Le pleurage est inférieur à 0,6 % de crête à crête sur une largeur de bande de 0,2 Hz à 20 kHz à 60 pouces/sec et la compensation de pleurage est prévue comme supplément facultatif.

Afin d'assurer la plus grande stabilité mécanique possible, les assemblages de tête, les guides de bande et le tourrévolver sont tous montés sur une seule plaque en aluminium anti-contrainte d'une épaisseur de 3,75 cm et la surface polie et rodée de cette plaque, plate jusqu'à un degré de tolérance de 5 dixièmes d'un millième de pouce, s'adapte aux surfaces finement polies des assemblages de tête. La largeur, l'espacement, le nombre et les tolérances de toutes les pistes sont conformes aux normes de l'IRIG.

Le magnétophone PI-2100 peut être utilisé à des températures de 40° F à 120° F et dans des conditions d'humidité relative non condensatrices allant jusqu'à 95 %. Il est muni de dispositifs antichocs et anti-vibrations répondant à toutes les spécifications militaires.

EE 80 753 *pour plus amples renseignements*

COMMUTATEUR D'ECHANTILLONNAGE ROTATIF

I.D.M. Electronics Ltd. 1 Arkwright Road, Reading, Berkshire

Ce commutateur d'échantillonnage rotatif à action rapide a été étudié par la société I.D.M. Electronics Ltd pour répondre à la plupart des besoins des systèmes télécommandés. Il peut être utilisé pour résoudre bon nombre des problèmes de la télémétrie, tels que l'interruption des signaux de courant continu, la production de formes d'ondes, la stabilisation d'amplificateurs, l'échantillonage à faible niveau de signaux d'extensimètres et de thermocouples, le comptage numérique ainsi qua d'autres applications à action lente ou rapide d'échantillonnage d'informations.

La combinaison de matériaux de contact en métal noble, de matériaux de raclage et de matériaux d'isolement a été effectuée afin de produire des conditions maxima de bruit électrique très faible, de force électromagnétique thermique réduite, de couple d'entraînement réduit, d'isolement élevé et de longue durée. Sur le plan mécanique, le commutateur comporte des roulements à billes de précision préchargés, lui assurant un comportement égal durant toute sa durée de fonctionnement. Sa construction particulière en fait un des plus petits commutateurs du genre et les constructeurs se sont attachés en particulier à étendre sa durée de vie sans entretien et sans l'emploi de bains d'huile.

La précision de la position des impulsions est d'un très haut degré, permettant ainsi d'utiliser les éléments sans matériel de démultiplexage automatique. Dans les systèmes à haute fréquence où il est difficile d'éliminer l'inter-couplage entre les canaux, ce type de commutateur fournit le couplage minimum possible entre segments.

La conception du commutateur permet une gamme étendue de configurations de sorte que le produit du nombre de pôles multiplié par le nombre de directions égale 24 (c'est à dire: 1 pôle 24 directions, 2 pôles 12 directions, etc.). Le commutateur peut être actionné par un moteur miniature à courant continu ou à courant alternatif et, étant donné que le banc du commutateur n'exige qu'un couple de quelques grammes/centimètre, même à une vitesse élevée, la consommation électrique est fort basse.

Les segments peuvent être reliés entre eux da manière à fournir une grande variété de rapports marque/espace. Des configurations de piste spéciales peuvent être étudiées sur demande. Les fils de conducteurs livrables sont soit simples soit à code de couleurs, soit encore blindés. Dans chacun des cas ils sont laissés libres pour pouvoir être fixés à la barrette à bornes du client.

EE 80 754 *pour plus amples renseignements*

SONOMÈTRE

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(*Illustration à la page 269*)

Le sonomètre Philips type PM6400 a été réalisé par The M.E.L. Equipment Co. Ltd pour servir d'instrument de base aux ingénieurs, architectes, travailleurs médicaux et autres s'occupant de mesure acoustique et du contrôle du bruit.

C'est un appareil entièrement transistorisé et il est très compact, son poids étant inférieur à 3 kg. Cette portabilité et le fait que l'appareil fonctionne sur batteries intérieures facilite grandement la mesure tant en laboratoire que sur place.

Une des extrémités de l'appareil est muni d'un microphone à cadre mobile omnidirectionnel et le niveau d'intensité du son est indiqué sur un instrument de mesure à cadre mobile étalonné de −7 à +10 dB. Des prises sont prévues pour pouvoir y brancher un enregistreur ou un filtre.

L'instrument est conforme aux normes de l'I.E.C. et couvre la gamme de 33 à 130 dB. Sa précision à 1 kHz, lorsqu'il se trouve à angle droit par rapport à la source acoustique, est de ±1 dB. La gamme de fréquences s'étend de 25 Hz à 15 kHz. Des réseaux A, B et C de type standard I.E.C. sont incorporés.

Le courant est fourni par deux cellules en nickel-cadmium qui assurent 20 heures de fonctionnement avant de devoir être chargé. Un chargeur de batterie type P.M.9000 est également fourni.

Le sonomètre mesure: 33 cm × 11,4 cm × 8,2 cm. Une valise portable de cuir est fournie avec l'appareil.

EE 80 755 *pour plus amples renseignements*

GÉNÉRATEUR DE SIGNAUX HAUTE-FRÉQUENCE

K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(*Illustration à la page 269*)

Le générateur de signaux G.30 est un appareil très pratique et peu coûteux couvrant les fréquences de 120 kHz à 240 MHz en sept bandes à commutation. Cette gamme étendue le rend utile pour la plupart des travaux expérimentaux et d'alignement. La sortie hautefréquence qui dépasse 100 mV peut être modulée intérieurement ou extérieurement, la modulation interne étant fournie à 400 Hz. La profondeur de modulation est réglable au moyen d'une commande de 0 à 50%. La fréquence de modulation de 400 Hz peut être également obtenue aux deux douilles du panneau frontal.

EE 80 756 *pour plus amples renseignements*

CONTRÔLEURS DE DÉCLENCHEMENT À THYRISTORS

Standard Telephones & Cables Ltd. Rectifier Equipment Division. Edinburgh Way, Harlow, Essex

(*Illustration à la page 270*)

Les contrôleurs de déclenchement à thyristors monophasés et triphasés que fournit STC assurent une impulsion d'amorçage à retard variable aux dispositifs de déclenchement des thyristors pour les applications telles que la régulation de l'alimentation, le contrôle de la vitesse de moteurs à courant continu et le contrôle de courant ou de tension d'une variété de circuits. Les contrôleurs ont une puissance nominale étendue et ils ont été étudiés pour de très longues durées d'utilisation sans la moindre panne.

Les contrôleurs de déclenchement à thyristors STC sont en effet des instruments électroniques d'une grande souplesse d'emploi, fournis sous forme d'éléments compacts pouvant être montés au niveau de la face du matériel à redresseurs au silicium. Un autre type d'assemblage permet le montage des éléments de déclenchement sur châssis à l'intérieur de postes de commande. Les connexions s'effectuent au moyen de fiches et de douilles ou au moyen de barrettes à bornes.

Les principales caractéristiques de ces contrôleurs sont:

(1) Alimentation autonome; seule une entrée de courant alternatif de secteur étant requise.
(2) Temps de réponse rapide.
(3) Signal d'amorçage à onde carrée donnant un contrôle complet pour les circuits inductifs.
(4) Hauteur d'impulsion insensible aux changements de tension secteur de +20 % et −50 %.
(5) Commande de thyristors jusqu'à 2 % de la tension de sortie maxima.
(6) Utilisation assurée avec toute les connexions normales de redresseurs.

EE 80 757 *pour plus amples renseignements*

COMMUTATEUR À LAME VIBRANTE

B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustration à la page 270)

Le commutateur à lame vibrante Gordos, type MR201, que vient de réaliser la société B & R Relays Ltd, est un commutateur à polarisation mécanique. Sa longueur hors tout est de 78,5 mm, la longueur du verre étant de 30,5 mm et le diamètre du verre de 5,1 mm. Le contact nominal est de 10 VA à un maximum de 0,25 A. Un nouveau matériau de contact, le "Di-met," est utilisé sur le commutateur MR201. "Di-met" est une marque déposée et donne des contacts à revêtement électrolytique de métaux dissemblables. L'or lui assure la stabilité, le cuivre la faible résistance et les autres alliages la dureté et la durabilité. Le nouveau matériau de contact se caractérise par sa très faible résistance de contact et son adaptabilité aux lampes de tungstène et autres charges semblables. Il est d'une plus grande stabilité inhérante que les matériaux de contact tels que le rhodium et l'argent. L'épaisseur du revêtement est réduite au minimum mais permet cependant le contrôle du différentiel de fonctionnement. En apparence, une des lames est de couleur argentée et l'autre de couleur cuivrée.

La gamme entière des commutateurs à lame vibrante Gordos peut être fournie soit avec contact en "Dri-met" ou en or diffus.

Le nouveau commutateur MR201 peut, entre autres, être incorporé au relais à lame vibrante B et R RO3 et RO4.

EE 80 758 *pour plus amples renseignements*

TRANSFORMATEURS MINIATURE

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

(Illustration à la page 270)

La société Parmeko Ltd fournit une gamme étendue de transformateurs et de selfs miniature pour transformateurs d'impulsions, acoustiques, de puissance et d'instruments, ainsi que pour inducteurs et selfs à tolérance réduite.

Cette gamme de transformateurs, série 4000, a reçu des certificats de haute qualité quant à la résistance à l'humidité et la température du Joint Service Radio Components Standardisation Committee. La construction générale de ces transformateurs et selfs est conforme à la spécification DEF 5214.

Les modèles ouverts de la série 4000 utilisent la gamme complète Parmeko de bobines simples et doubles et ils sont particulièrement indiqués pour les applications exigeant un degré élevé de compacité. La plupart de ces modèles ouverts sont munis de noyaux en ferrite. Les terminaisons sont effectuées sur des broches moulées dans les joues des bobines. Trois types de griffes peuvent être fournis pour différents montages.

Trois formes différentes d'assemblages sont prévues: encapsulée, en coffret d'aluminium, et à blindage magnétique. Les modèles encapsulés sont sous moulage de résine et ont l'avantage supplémentaire d'une rigidité mécanique et d'un isolement électrique accrus. Les modèles sous moulage et dans des coffrets en aluminium emploient la résine d'époxyde ou tout autre matière appropriée pour présenter une forme simple de montage avec tous les avantages des modèles encapsulés. Les problèmes d'interférence magnétique sont résolus par l'emploi de coffrets en mumétal, et des modèles à double ou triple blindage peuvent être fournis sur demande. Les terminaisons de ces différents modèles peuvent être fournis pour broches appropriées ou pour câbles mobiles, selon la forme de l'assemblage et le nombre d'extrémités d'enroulement.

EE 80 759 *pour plus amples renseignements*

AMPLIFICATEURS POUR MOTEURS À CIRCUIT IMPRIMÉ

Pye Printed Motors Ltd, St. Andrews Road, Cambridge

(Illustration à la page 270)

Le moteur à circuit imprimé, en raison de sa conception, pose des problèmes dans l'étude d'amplificateurs d'asservissement appropriés. Le moteur étant essentiellement un dispositif à faible tension et à courant élevé, l'amplificateur d'asservissement doit être conçu en fonction de ces caractéristiques si l'on veut en tirer tous les avantages possibles.

La société Pye Printed Motors Ltd a maintenant mis au point une chaîne d'amplification se composant d'un amplificateur de puissance avec son dispositif d'entraînement et son amplificateur opérationnel permettant de contrôler les moteurs de la série 4 dans toutes les conditions normales et les moteurs plus grands avec une certaine limite à la vitesse maxima (en raison de leur f.é.m. plus élevés).

Les caractéristiques essentielles de l'amplificateur de puissance PA.1G et de son amplificateur d'entraînement D.A.1G sont:

Alimentation
30 V c.c., à prises médianes (±15 V c.c.).
Courant continu
8,75 A max.
Gain de tension
D.A. 1G + P.A. 1G —2 fois.
Impédance de sortie nominale
0,05Ω
Dissipation maxima à une température ambiante de 25° C.
34 W.

Pour les applications exigeant des courants supérieurs à 8,75 A deux ou plusieurs étages d'amplificateurs de puissance peuvent être placés en parallèle et jusqu'à trois de ces étages peuvent être entraînés par un seul amplificateur d'entraînement D.A. 1G.

L'amplificateur opérationnel O.A. 1G est normalement utilisé avec réaction, choisie pour définir les caractéristiques de gain de phase du circuit d'asservissement. L'amplificateur a la spécification suivante:

Alimentation
24 V avec prise médiane (±12 V c.c.)
Consommation de courant au repos
4,5 mA
Gain du circuit ouvert nominal
60-80 dB
Stabilité avec réaction de 100 %
200 Hz
Largeur de bande du circuit ouvert
200 Hz
Dérive de tension d'entrée à une impédance de source de 1 kΩ
500 V/°C à 25° C.

Les amplificateurs opérationnels et d'entraînement sont des blocs modulaires mesurant environ 3,81 cm × 2,54 cm × 1,90 cm de hauteur et ils peuvent être montés sur plaquette de circuit imprimé. Les broches de connexion sont prévues pour un espacement de matrice de 0,25 cm.

L'amplificateur de puissance entoure une source froide à ailerons profilés. Ses dimensions hors tout approximatives sont: 15,24 cm × 10,79 cm × 7,62 cm.

EE 80 760 *pour plus amples renseignements*

ÉLÉMENTS SENSIBLES INDUCTIFS

Electronic Machine Control Ltd, Sherman Road, Bromley North, Kent

(Illustration à la page 271)

La société Electronic Machine Control Ltd vient de compléter sa gamme actuelle d'éléments sensibles photoélectriques à ultra sons et du type à capacitance par une nouvelle série d'éléments inductifs pour le contrôle électronique industriel et les systèmes de comptage.

Fournis dans une gamme étendue de type standard, les nouveaux dispositifs sont des éléments sensibles non-contactants, influencés par la présence d'objets métalliques ferreux ou non-ferreux. Ils sont de construction robuste, entièrement scellés et peuvent être utilisés dans l'eau, dans l'huile et dans les liquides non-corrosifs.

En raison de leur simplicité, puisqu'ils ne comportent pas de parties mobiles, ces éléments sensibles peuvent fonctionner pendant de longue périodes dans des conditions sévères de température et de vibrations et n'exigent aucun entretien en cours de service.

L'élément sensible mesure à peine 4,44 cm × 2,22 cm de diamètre et ses faibles dimensions permettent son installation dans des espaces très restreints où d'autres éléments sensibles logeraient avec difficulté. Toute masse métallique, grande ou petite, peut être détectée même à travers le verre ou les matières plastiques.

Un appareil électronique fonctionnant sur secteur et mesurant 12,06 cm × 19,05 cm × 5,08 cm est fourni avec chaque élément sensible et il peut être monté à une distance de 15 mètres de ce dernier, ou davantage. C'est un appareil entièrement transistorisé dans l'intérêt de sa sécurité de fonctionnement à long terme, il n'exige aucun temps de chauffage et, comme l'élément sensible lui même, aucun entretien. Une seule commande de sensibilité et un voyant lumineux sont fixés sur le couvercle frontal pour permettre la mise en route. Une fois cette dernière effectuée, nul réglage supplémentaire n'est nécessaire.

L'appareil électronique commande un relais interne comprenant deux paires de contact d'une valeur nominale de 250 V, 5 A, non-inductifs.

EE 80 761 *pour plus amples renseignements*

BROCHES DE BORNES

Vero Electronics Ltd, South Mill Road, Southampton, Hampshire

(Illustration à la page 271)

La société Vero Electronics Ltd a mis au point une nouvelle broche de borne, numéro 2141, pouvant être utilisée avec toute plaquette Veroboard (avec ou sans bandes de cuivre) dont le diamètre des ouvertures est de 1,3 mm, ou toute plaquette de borne ou de circuit imprimé perforé ou percé de trous du même diamètre. Les cannelures sous la plaquette assurent une fixation très ferme à la broche qui ne peut donc tomber de la plaquette pendant ou après le soudage. Ces nouvelles broches sont en laiton à revêtement d'étain pour faciliter le soudage.

EE 80 762 *pour plus amples renseignements*

PLAQUETTES DE COMPOSANTS À FICHES

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustration à la page 271)

La société Keyswitch Relays Ltd offre maintenant une plaquette pour composants à fiches, type 304 constituant un accessoire peu coûteux permettant aux réalisateurs de circuits d'utiliser un module commun pour relais, circuits constitués de corps solides et miniaturisés, petits moteurs et blocs d'alimentation.

La nouvelle plaquette comprend la base à fiches moulée à 22 directions du relais Keyswitch P33, avec une plaquette de composants de circuit enfermée dans une housse transparente anti-poussière.

Au cours des essais de production, et pour démontrer les possibilités de la nouvelle plaquette, la société Keyswitch a inseré avec succès dans un élément standard un étage de compteur annulaire complet avec bloc d'alimentation et comprenant 11 tubes de déclenchement, 11 diodes, 65 résistances et 35 condensateurs, tous fixés à l'intérieur d'un espace de montage de 79,4 × 56,3 × 28,1 mm. Le type 304 peut être fourni également sans plaquette de composants, si nécessaire.

Il se caractérise en particulier par la facilité d'entretien et de modification, car tous les éléments endommagés par une panne de circuit ou de composants peuvent être remplacés immédiatement à partir des pièces de rechange en stock. Des douilles supplémentaires peuvent être montées sur le matériel existant au cours de la fabrication initiale, permettant d'effectuer aisément et économiquement des modifications ultérieures.

Les douilles du panneau ne peuvent recevoir que les éléments pour lesquels elles sont prévues, et ce en raison d'un dispositif de manipulation non-interchangeable, tandis que les griffes de serrage en acier à ressort dans un choix de couleurs peuvent être fournies pour les besoins de codage et d'identification.

La puissance nominale de l'assemblage de fiches et de douilles (accouplées) est de 10 A, avec une résistance de contact de 5 mA et une tension de régime de 750 V c.a./c.c. maximum. La résistance d'isolement est supérieure à 100 MΩ à 500 V c.c.

La plaquette de composants est en papier laminé à agglomérant de résine synthétique, auquel sont agglomérées des bandes de cuivre longitudinales, de 0,4 mm d'épaisseur, à intervalle de 3,8 ou de 5,1 mm. Les perforations forment une matrice régulière de trous. La surface de bobinage est de 83 × 56,3 mm et la puissance nominale du courant (bande de cuivre) s'étend de 1 A, avec une montée de température de 1° C, à 5 A avec une montée de température de 39° C. La résistance est d'environ 0,045 Ω/pouces et l'isolement est supérieur à 100 MΩ. La tension de régime maxima est de 500 V c.c. entre bandes de cuivre adjacentes.

EE 80 763 *pour plus amples renseignements*

CONNECTEURS DE BORDS

Ferranti Ltd, Kings Cross Road, Dundee, Scotland

(Illustration à la page 271)

A l'instar de la gamme croissante de connecteurs électriques produits par la société Ferranti Ltd, les connecteurs de bords pour circuit imprimé se distinguent par leur degré élevé de fiabilité réalisé par l'emploi d'un ressort à contact à taux extrêmement réduit et des surfaces de contact dorées non poreuses.

Les griffes de contact des connecteurs de bords emploient des ressorts à taux réduit indépendants qui permettent de contrôler les forces de contact avec précision et dans des limites serrées durant la fabrication. Pinces et ressorts sont en bronze phosphoreux et les pinces ont un minimum de 0,005 mm d'or dur dans la surface de contact. La réalisation mécanique garantitle contact avec la plaquette de circuit dans toutes les conditions d'emploi et avec toutes les épaisseurs spécifiées de plaquette à condition d'employer la force voulue. Les pinces de contact sont en fil métallique laminé afin de ne pas risquer d'endommager les conducteurs de la plaquette et les surfaces de contact sont arrondies.

Les contacts ont été dorés selon des normes très rigoureuses. Des échantillons de chaque lot sont soumis au contrôle de galvanoplastie DEF5325. Ils sont également soumis à des essais de flexion et de brunissage. De plus, des sections sont prélevées de manière à pouvoir mesurer l'épaisseur du revêtement dans la zone de contact. Des méthodes statistiques de contrôle de qualité sont utilisées à tous les stades importants de la fabrication.

Les moulages des connecteurs sont soit en phthalate diallyle de verre réfractaire ou en matière prénolique de nylon de haute qualité. Le moulage en phthalate offre cependant une grande stabilité dimensionnelle aux température élevées. Des rayons aigus sont évités à tous les points où la plaquette entre en contact avec le moulage et il y a une entrée chanfreinée pour faciliter l'accès à la plaquette. Les clefs de polarisation peuvent être insérées dans n'importe laquelle des positions polaires sans enlever ou endommager le contact.

Ces connecteurs sont fournis dans les formats 8, 16, 24, 32 et 40 avec contacts simples ou bilatéraux munis de chevilles. La capacité de courant est de 1 A par pôle (à 100° C) et la tension continue de régime est de 450 V. En cas d'utilisation avec une plaquette de circuit imprimé non chanfreiné, la force d'insertion est d'environ 15,6 g par contact, mais si on emploie une plaquette chanfreinée, la force d'insertion est d'environ 100 g. La force totale d'insertion est directement proportionnelle au nombre de pôles.

La pression de contact contrôlée est suffisante pour assurer une résistance de contact réduite, inférieure à 3 mΩ, même aux tensions de régime les plus faibles.

EE 80 764 *pour plus amples renseignements*

JAUGE DE TEMPÉRATURE

Startronic Ltd, 117a Malden Road, New Malden, Surrey

(Illustration à la page 272)

Il s'agit ici d'un instrument compact et portatif fonctionnant sur batterie pour mesurer la température de $-20°$ C à $+120°$ C en deux gammes, $-20°$ C à $+55°$ C et $+50°$ C à $+120°$ C avec une précision de $\pm1°$ C.

Il utilise des thermistors pleinement interchangeables pour la perception de température.

Les échelles ont une excellente linéarité et l'instrument de mesure fournit une longueur d'échelle de 9,52 cm.

L'instrument peut être placé à une distance considérable de la sonde sans en affecter sérieusement la précision. Par exemple, un conducteur d'une résistance de 2 Ω ne donne qu'une erreur de $\frac{1}{4}°$ C à 100° C.

Des thermistors spécialement adaptés l'un à l'autre permettent à l'utilisateur de les fixer au points voulus dans le matériel de mesure des températures de surface. Des sondes peuvent être fournies pour mesurer les températures de gaz ou de liquides.

De plus, un sélecteur à dix directions et divers câbles d'extension sont prévus.

EE 80 765 *pour plus amples renseignements*

COMPTEUR-MINUTERIE

J.A.C. Electronics Ltd, Station Estate, Blackwater, Camberley, Surrey

(Illustration à la page 272)

La société J.A.C. Electronics Ltd produit maintenant un nouvel instrument transistorisé à fréquence numérique et à intervalles de temps, comportant une base de temps pilotée au quartz et à variation complète, pour assurer une lecture directe dans tous les éléments prédéterminés (par exemple, tours/minute). C'est un appareil à impédance d'entrée élevée et à affichage en ligne d'une lecture facile. Il comprend un indicateur donnant une lecture efficace à 7 chiffres pour le prix d'un indicateur à 4 chiffres.

Des périodes de comptage entre 1 msec et 10 sec en plots de millisecondes sont prévues pour la mesure des fréquence de 0,1 Hz à 300 kHz; les intervalles de comptage peuvent être contrôlés manuellement ou extérieurement par signaux ou par fermeture de contact pour la détermination générale du nombre de phénomènes par temps unitaire.

Les sondes magnétiques ou photoélectriques (avec lampe intégrée) sont destinées aux applications tachymétriques.

EE 80 766 *pour plus amples renseignements*

MINUTERIES DE PROCESSUS

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustration à la page 272)

La société Kent Precision Electronics a mis au point une nouvelle gamme de minuteries de processus constitués de corps solides pour les applications industrielles. Ces instruments font partie de leur gamme d'instruments de contrôle constitués de corps solides. Ils sont prévus pour le minutage précis de la puissance électronique dans la gamme de 10 msec à 15 sec avec une précision totale de ±1 %.

Des instruments simples ou doubles qui actionnent des relais électromagnétiques standard sont logés dans le coffret à instruments pour montage sur panneau standard comme on peut le voir dans notre gravure. Il existe également des instruments avec sorties par impulsions pour amorcer des circuits de thyristors pour le contrôle de la puissance jusqu'aux valeurs nominales de thyristors maxima.

Les instruments se composent essentiellement d'un circuit de minutage RC et d'un détecteur précis de seuil de tension à transistors qui entrent en action lorsque la tension dans le circuit de minutage atteint une valeur déterminée et amorce la sortie de thyristors appropriée. Pour la mise en oeuvre du relais, un thyristor de faible puissance est utilisé pour contrôler un relais standard. Pour le contrôle des corps solides l'impulsion de sortie est utilisée pour déclencher les thyristors appropriés.

Une des caractéristiques les plus intéressantes de cette gamme est la présence d'un instrument double à l'intérieur du coffret standard pour montage sur panneau. Cette double caractéristique permet à l'instrument d'être interconnecté, de sorte qu'après la première période de temps écoulé, il se produit une deuxième période, l'ensemble du processus pouvant être remis en cycle avec des rapports réglables d'arrêt/marche. En variante, l'instrument peut être utilisé sous forme de deux minuteries séparées avec commandes concentriques.

EE 80 767 *pour plus amples renseignements*

AMPLIFICATEUR SÉLECTEUR DE FRÉQUENCE

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Illustration à la page 273)

La société H. Tinsley è Co. Ltd vient d'introduire un nouvel amplificateur sélecteur de fréquence transistorisé. Il s'agit d'un instrument portatif et fonctionnant sur batterie ayant de nombreuses applications. Il peut, par exemple, être utilisé avantageusement avec des ponts ou des diviseurs de tension inductifs pour effectuer des mesures à un certain nombre de fréquences. La portabilité de l'appareil a été facilitée par sa compacité et son poids réduit. Il mesure 30,48 cm × 10,16 cm × 21,59 cm et son poids est d'environ 2,25 kg.

Il peut détecter des signaux d'entrée de 0,1 μV à 10 V et sa fréquence peut être accordée dans quatre gammes de 10 Hz à 100 kHz. La fréquence voulue s'obtient sur trois cadrans à décades réglables à une précision de 1 % de la fréquence jusqu'à 10 kHz, avec un facteur Q supérieur à 30 et une sensibilité telle que 0,1 μV peut être détecté sur toute la gamme.

Trois étages d'amplification sont incorporés à l'appareil. Ils utilisent des transistors à gain élevé et à faible bruit. Le gain de tension de l'amplificateur est de l'ordre de 10^6. Ce gain ne dépend pas de la température grâce à l'emploi de transistors au silicium à faible perte. L'impedance d'entrée de l'instrument varie en fonction du degré d'atténuation. L'atténuateur est divisé en 11 plots de 10 à 120 dB. Pour les signaux d'entrée supérieurs à 10 μV, l'impédance d'entrée est de 10 kΩ avec 10 à 50 dB d'atténuation, montant à 1 MΩ avec une atténuation de 50 à 120 dB pour les entrées plus importantes.

Il existe également un autre modèle, le type 5711, étalonné en fréquence angulaire.

EE 80 768 *pour plus amples renseignements*

JAUGE D'ÉPAISSEUR PORTATIVE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 273)

La plus récente des jauges d'épaisseur à ultrasons réalisée par Dawe Instruments Ltd porte le nom de Sonoray 30, type 1805A. Bien que principalement conçue pour le fonctionnement sur batterie en campagne, sa précision et sa résolution sont égales à celles des grandes jauges pour chaînes de montage.

Les travaux de jauge par ultrasons ont été limités dans le passé par la nécessité de devoir employer du courant de secteur, qu'il n'était pas toujours possible d'obtenir et qui ne peut, en tous cas, être fourni de manière pratique par un câble mobile lorsque l'opérateur doit se frayer un chemin sur des escaliers, des échaffau dages et d'autres obstacles de ce genre. L'appareil type 1805 a été étudié pour pouvoir surmonter cette difficulté. C'est un instrument portatif mesurant 38,11 cm × 24,13 cm × 21,43 cm de haut et pesant 7,25 kg avec tous ses accessoires et normalement mis en oeuvre par sa propre batterie rechargeable. Il peut cependant être actionné par courant secteur et utilisé tant pour la jauge d'épaisseur que pour celle des pailles.

Le générateur produit des impulsions de 1,5 à 5 MHz, dont les réflexions sont enregistrées sur un écran de 6,35 cm × 6,35 cm. L'échelle peut être agrandie jusqu'à ce que sa largeur totale représente une épaisseur d'acier de 1,27 cm. La précision de jauge est de $\pm0{,}005$ pouce pour des épaisseurs de 0,25 cm à 15,24 cm. Pour le détection des pailles, exigeant un degré de précision moins critique, la gamme s'étend jusqu'à 254 cm.

Les transducteurs doivent être adaptés à la fréquence utilisée et ils sont normalement du type piézoélectrique Z à performance élevée. L'instrument standard est fourni avec un câble de transducteur de 182 cm, mais des câbles d'une longueur de 60,96 m peuvent être employés sans modifications. L'appareil permet les essais par faisceaux droits et angulaires, par transmission directe et par immersion.

EE 80 769 *pour plus amples renseignements*

INSTRUMENT DE NETTOYAGE À ULTRASONS

Sonics Division, Elliott-Automation Ltd, Bath Road, Beenham, Reading, Berkshire

(*Illustration à la page 273*)

Le nouvel instrument de nettoyage à ultrasons mis au point par la division d'acoustique de la société Elliott-Automation Ltd est destiné en particulier au nettoyage de composants délicats dans les laboratoires et l'industrie de précision, ainsi qu'aux travaux d'horlogerie et de bijouterie.

C'est un appareil extrêmement sûr et entièrement autonome. Tant le réservoir d'une contenance de 2 litres (mesurant 15,24 cm × 11,43 cm × 12,7 cm intérieurement) que le générateur de 60 W sont logés dans un robuste coffret mesurant 35,56 cm × 20,32 cm × 24,3 cm. Le commutateur arrêt/marche à bouton-poussoir est la seule commande nécessaire car l'accord a été éliminé par l'emploi de circuits transistorisés en liaison avec le transducteur "Multipower" Elliott-Acoustica. L'efficacité de nettoyage maxima est toujours assurée quel que soit le contenu du réservoir ou les changements de température. Un robinet de vidange permet de vider le réservoir. Une connexion de tuyaux sur le robinet permet d'utiliser une pompe de circulation et un filtre.

L'appareil peut être utilisé par des opérateurs non-spécialisés et la stabilité à long terme de la sortie de puissance permet aux temps de nettoyage d'être d'une durée standard.

EE 80 770 *pour plus amples renseignements*

MÉTALLOSCOPE À ULTRASONS

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(*Illustration à la page 273*)

Le détecteur de pailles à ultrasones, Attenutector Mk. IV, est un appareil perfectionné dont la mise au point vient d'être achevée. Il a été conçu pour le contrôle non-destructeur de matières plastiques et de graphite mais il peut également être utilisé pour une variété d'autres applications telles que la détection de bulles dans les liquides, la vérification de la liaison d'une matière plastique à une autre matière plastique ou d'une matière plastique au métal, ou encore du métal au métal, et enfin de la variation de densité dans les liquides.

Cet instrument, comme son nom l'indique, fonctionne suivant un principe de transmission directe, dont le rapport d'atténuation à travers la pièce soumise à l'essai, en fonction du signal reçu, peut se réduire à 1 000 000 : 1. Par conséquent les technique d'immersion sont prescrites pour éviter toute variation de contact entre les sondes et la pièce soumise à l'essai.

Des sondes de type standard sont fournies pour les fréquences de 250 à 650 kHz et d'autres fréquences jusqu'à 2 MHz peuvent être prévues sur commande spéciale.

Le système est en général plus sensible que le contrôle par rayons X et il coûte beaucoup moins cher, tant en ce que concerne les frais initiaux que les frais d'utilisation.

EE 80 771 *pour plus amples renseignements*

CONVERTISSEUR ANALOGIQUE-NUMÉRIQUE

Moore, Reed & Co. Ltd, Woodman Works, Durnsford Road, London, S.W.19

(*Illustration à la page 273*)

La société Moore Reed & Co. Ltd vient d'ajouter à sa gamme d'appareils de contrôle un convertisseur électromécanique dans lequel un plot d'un dispositif d'entraînement produit un changement de chiffre sur une sortie de codage.

Les applications comprennent la conversion de signaux de courant alternatif biphasé ou triphasé en codes numériques, binaires ou de Gray. Lorsque l'appareil est utilisé avec un élément de soustraction, il peut produire une sortie analogique synchronisée, de résolution ou potentiométrique, à partir d'une entrée numérique. Le convertisseur répond à un train d'impulsions d'un élément de programmation, injecté, à travers un circuit logique transistorisé, au moteur entraînant l'encodeur, donnant ainsi des signaux numériques ou analogiques simultanés pour le contrôle ou l'affichage.

L'emploi d'engrenages de précision empêchant tout jeu nuisible permet de réduire les erreurs à environ 5 minutes d'arc.

Les dimensions extérieures de l'appareil sont d'environ 12,7 cm × 3,81 cm × 6,98 cm. L'appareil peut être prévu pour le balayage binaire en V, ainsi que pour des sorties numériques du code Gray ou Gilman.

EE 80 772 *pour plus amples renseignements*

Résumés des Principaux Articles

Calculatrice de la fonction de transfert par R. W. Levinge

Résumé de l'article aux pages 218 à 224

Le comportement d'un circuit linéaire à quatre bornes ou d'un système de contrôle est défini en toutes circonstances par sa fonction de transfert. Un nouvel instrument a ét réalisé qui permet, par la mesure directe à travers le réseau ou système, de calculer sa fonction de transfert par la variable complexe p. La méthode d'analyse pratiquée jusqu'ici était basée sur la mesure des phases et du gain ou de la réponse transitoire. L'emploi de la calculatrice de la fonction de transfert obvie à la nécessité du travail considérable qu'implique le calcul de la fonction de transfert d'après ces données.

A condition de connaître tous les paramètres, on peut calculer aisément la fonction de transfert d'un système. L'appareil fournit donc le moyen de mesurer la fonction de transfert effective du système et de vérifier les paramètres individuels par comparaison. En variante, si un système ou un réseau comportent une fonction de transfert déterminée, les termes de cette dernière peuvent être établis sur la calculatrice. Les commandes préréglées du système peuvent être fixées sur spécification. La calculatrice peut être utilisée également comme appareil de contrôle pour accepter ou rejeter le système.

Échelle de comptage à diode tunnel à action rapide par M. N. S. Swamy

Résumé de l'article aux pages 225 à 229

Cet article décrit en détail la conception et le fonctionnement d'une échelle de comptage de décades à diodes tunnel. Les tolérances étendues de cette échelle sont prévues pour les courants de polarisation continue ainsi que pour les amplitudes d'impulsions d'entrée. L'échelle elle-même peut fonctionner jusqu'à une fréquence de 25MHz, cependant que le circuit d'entrée peut recevoir jusqu'à 10MHz, L'échelle de comptage de décades peut fournir des impulsions négatives ou positives d'une amplitude de 0,5V.

Intégrateur transistorisé par J. A. Effenberger et R. Steele

Résumé de l'article aux pages 230 à 234

Cet article traite d'un intégrateur de déclenchement produisant une tension de rampe positive ou négative. Les auteurs présentent à ce sujet un circuit pratique entièrement transistorisé consistant en un amplificateur à couplage direct relié, par une résistance normale et un réseau de réaction de capacité, à un commutateur électronique de déclenchement pour décharger la capacité. Durant la période de retour, la tension de sortie de crête maxima de 28V est déchargée à 5% près après environ 50µS puis à moins de 10mV près après 60µS supplémentaires. La linéarité de la tension de sortie de rampe est supérieure à 2.10^{-4}.

Un contrôleur de transistors par M. M. Winn

Résumé de l'article aux pages 234 à 235

Cet appareil de contrôle de transistors de conception simple permet de vérifier le comportement des transistors à un courant au collecteur choisi par l'opérateur. Le résultat de ce contrôle est indiqué directement sur un appareil de mesure. Le processus de contrôle ressemble dans une grande mesure au fonctionnement d'un transistor dans un circuit de commutation.

Amplificateur de courant électronique pour système d'analyse à microsonde électronique par J. B. Davey

Résumé de l'article aux pages 236 à 239

Les facteurs fondamentaux déterminant le rapport signal/bruit d'un amplificateur à faibles courants électroniques secondaires sont examinés dans cet article qui montre que, dans certaines circonstances, on ne trouve aucun avantage important dans l'emploi de la multiplication électronique avant l'amplification.

L'article donne un schéma de réalisation pratique pour un amplificateur de courants allant jusqu'à 10^{-10}A à des largeurs de bande variant du courant continu jusqu'à 10kHz. Le comportement de cet amplificateur est également décrit.

Diviseur de fréquences microélectronique à rapport de division variable par B. Preston

Résumé de l'article aux pages 240 à 244

L'auteur décrit le fonctionnement d'un diviseur de fréquence permettant de fixer le rapport de division à n'importe quel nombre entier voulu. Il décrit également en détail le circuit de l'appareil et explique les méthodes employées pour réaliser ce circuit sous forme microélectronique. Les résultats et les formes d'ondes obtenue de ce circuit sont indiqués.

Un nouveau synthétiseur de Fourier basé sur le multivibrateur accouplé à l'émetteur par A. D. Bond

Résumé de l'article aux pages 245 à 247

Cet instrument produit une forme d'onde fondamentale sinusoidale à fréquence fixe et des harmoniques jusqu'au dixième. Toutes les harmoniques sont à variation continue en amplitude et en phase et elles sont également sommables pour synthétiser une variété de formes d'ondes.

Magnétomètre de générateur par P. I. Ghelfan

Résumé de l'article aux pages 248 à 251

Cet article décrit un magnétomètre de générateur à lecture directe à haute sensibilité et d'un maniement très simple pour la mesure de faibles facteurs de magnétisation résiduelle.

Un dispositif de compensation électronique annule l'effet magnétique erroné ne se rapportant pas à l'échantillon. Afin d'éviter tout problème de protection, des circuits RC sont exclusivement utilisés dans ce magnétomètre.

Générateur d'impulsions à haute tension par I. M. Paz

Résumé de l'article aux pages 252 à 255

L'auteur décrit un générateur d'impulsions à haute tension fournissant, sur charge capacitive, des impulsions d'amplitude variables: (500-1000V), largeur: (10-110μs) et taux de fréquence d'impulsions: (0-1.000 impulsions par seconde).

Les changements de taux ou de tension aux extrémités de tête et d'accompagnement des impulsions sont inférieurs à 1μs par kV.

La commutation s'effectue directement sur la ligne de haute tension, la capacitance de charge étant chargée durant le temps de montée et déchargée durant le temps de chute par les éléments de commutation.

L'instrument emploie comme éléments de commutation principaux des thyratrons à l'hydrogène miniature du type à polarisation nulle.

Circuits d'amorçage perfectionnés à redresseurs pilotés au silicium par D. Ciscato et L. Mariani

Résumé de l'article aux pages 256 à 258

L'emploi accru des redresseurs pilotés au silicium pour la commande et la transformation de puissance a entraîné l'étude et l'application d'un grand nombre de dispositifs d'amorçage.

Cet article présente deux de ces circuits constitués de corps solides, dont l'un est prévu pour des charges inductives monophasées et l'autre pour les ponts contrôles triphasés.

Un diviseur analogique transistorisé par G. Gooberman et B. Jones

Résumé de l'article aux pages 259 à 261

Les auteurs décrivent dans cet article le schéma de réalisation d'un diviseur analogique transistorisé. Ce diviseur peut donner une sortie prévisible à plus de 0,5% et une linéarité supérieure à ±0,2% pour des tensions variant entre 0,7 et 20 volts. Ce diviseur utilise une méthode à impulsions d'espacement par marque variable dont la diode tunnel forme le coeur du générateur d'impulsions.

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte **auf Grund der von Herstellern gemachten** Angaben.

Übersetzung der Seiten 268 bis 273

Doppelimpulsgeber

Vertrieb: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1

(Abbildung Seite 268)

Die E-H Research Laboratories in Oakland, Kalifornien, produzieren zur Zeit ihren Impulsgenerator 132L in England, und zwar werden positive oder negative Einzel- oder Doppelimpulse mit einer Spitzenamplitude von 50 V an einen Abschluss von 50 Ω und mit Anstiegs- und Abfallzeiten von 10 ns erzeugt.

Das Modell 132L kann entweder getriggert oder freilaufend betrieben werden, und zwar zwischen 10 Hz und 3,5 MHz als Einzelimpulsgeber, oder zwischen 10 Hz und 1 MHz als Doppelimpulsgeber. Gegen Aufschlag ist eine Abwandlung lieferbar, die Betrieb bis zu 1 Hz herunter gestattet.

Impulsdauer und -verzögerung sind durch Frontplattenelemente kontinuierlich zwischen 100 ns und 10 ms regelbar.

Komplexe Impulsreihen lassen sich mit Hilfe von internen Torschaltungen erzeugen: Anlegen einer Gleichspannung von −5 V an eine Frontplattenbuchse sperrt die Ausgangsimpulse, solange sie angelegt bleibt.

Die Zeitschaltung des 132L ist volltransistorisiert. Eine wesentliche Eigenschaft der in der Konstruktion des Gerätes angewandten Schaltung ist, dass keine voreingestellte Wellenformjustierung erforderlich ist: die Ausgangsimpulse sind ungewöhnlich sauber und schnell und ohne Wechselwirkung zwischen Amplitude, Frequenz, Dauer und Verzögerungsregelung.

EE 80 751 *für weitere Einzelheiten*

Automatischer Taupunktanzeiger

Shaw Moisture Meters, Rawson Road, Westgate, Bradford, 7, Yorkshire

(Abbildung Seite 268)

Genaue absolute Anzeigen der Taupunkttemperatur von Luft oder Gas gibt ein mit "Thermodew" bezeichnetes Instrument, das die neuen thermisch gekühlten Halbleiter, die modernste Fotozelle und Thermistor-Temperaturmessung zur Schaffung eines in sich geschlossenen, automatischen, thermisch gekühlten Taupunktmessgerätes anwendet. Eine Ausführung für Registrierung ist auch lieferbar.

Die Genauigkeit der Taupunktbestimmung ist durch die Konstruktion bedingt. Das Licht einer von einer Konstantspannungsquelle gespeisten Lampe wird auf eine Fotozelle reflektiert, wenn sich Tau auf einer Spiegelfläche bildet. Der zur Kühlung des Spiegels durch den Thermobaustein fliessende Strom wird dann unterbrochen, so dass die Spiegelfläche, deren Temperatur gemessen wird, dauernd zwischen dem betauten und nicht betauten Zustand pendelt. Dadurch wird verhindert, dass der Spiegel durch Drift in einer der Schaltungen unter die Taupunkttemperatur gekühlt wird. Eine Signallampe auf der Frontplatte bestätigt, dass der Tau-kein Tau-Zustand aufrechterhalten bleibt. Diese Arbeitsweise wird als wesentlicher Fortschritt gegenüber den früheren bei der automatischen Bestimmung der Taupunkttemperatur verwendeten Servosystemen angesehen. Es werden durchgehend Festkörper-Bauelemente benutzt.

Das Standardmodell hat eine Uhr mit einstellbarem Einschaltkontakt und regelbarem Intervall zwischen Betriebsperioden, normalerweise eine Minute in jeder 10-Minuten-Periode. Will man eine Taupunktbestimmung, ehe die Uhr den fälligen Kühlungsvorgang einschaltet, so kann ein vorgespannter Kippschalter (markiert Auto. aus—ein) manuell heruntergedrückt werden, bis sich auf der Skala eine stetige Taupunktanzeige einstellt. Dieses Verfahren wurde gewählt, um dem "Thermodew" eine zuverlässige Betriebsweise zu geben und zu verhindern, dass das Gerät dauernd eingeschaltet bleibt, auch wenn es nicht gebraucht wird. Mehrjähriges fehlerfreies Arbeiten wird erwartet.

Hinten ist eine Luftpumpe angebaut, so dass der Taupunkt der Umgebungstemperatur gemessen werden kann. Die ebenfalls von der Schaltuhr gesteuerte Pumpe bläst einmal jede 10 Minuten durch die Messkammer. Sie saugt Luft durch ein hinten angeordnetes Filter aus gesinterter Bronze an. Wenn der Taupunkt zur Prüfung zugeführten Gases benötigt wird, entfernt man den Schlauch vom Einlassstutzen. Die Messkammer hat Einlass- und Auslassstutzen für Gas, die mit der Gasleitung verbunden werden können. Die Gasleitung soll unter leichtem Druck stehen, da die Pumpe nicht in dieser Leitung liegt und eine geringe Durchströmung aufrechterhalten werden muss.

EE 80 752 *für weitere Einzelheiten*

Instrumentierungs-Magnetbandgerät

Vertrieb: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Abbildung Seite 269)

Das Hauptmerkmal des Instrumentierungs-Magnetbandgerätes PI-2100 der Precision Instrument Co. Ltd ist seine Fähigkeit, den Vorschriften eines Pflichtenblattes zu genügen, das normalweise nur von viel grösseren und aufwendigeren Labor-Ausrüstungen erwartet wird. Die Leistungsfähigkeit ist tatsächlich so hoch, dass man das Gerät für 99 Prozent der zu erwartenden Datenaufzeichnungsaufgaben einsetzen kann; hinzu kommen noch die Vorteile kleiner Abmessungen und Tragbarkeit, die Aussendiensteinsatz unter sonst unmöglichen Bedingungen ermöglichen. Weiterhin kann man noch über Entfernungen bis zu 30 m fernsteuern und auch das Gerät aus den verschiedensten Stromversorgungen speisen, u.a. Batterien, 28 V Gleichstrom und Wechselstromquellen bei verschiedenen Frequenzen. Ein weiteres Merkmal, das zum reibungslosen Arbeitsablauf beiträgt, ist die speziell entwickelte Bandkassette, die Auswechseln des Magnetbandes einschliesslich automatischem Einlegen des Bandes über die Köpfe in wenigen Sekunden möglich macht. Eine Endloskassette mit 730 cm Bandlänge ist gleichfalls lieferbar.

Das Gerät kommt in zwei Ausführungen, als 7-Kanal- oder 14-Kanal-Modell, und jedes Modell ist mit vier oder sechs Bandgeschwindigkeiten lieferbar. In der Ausführung mit vier Geschwindigkeiten gibt es 152, 76, 38 und 19 cm/s, in der mit sechs Geschwindigkeiten 152, 76, 38, 19, 9,5 und 4,75 cm/s. Eine fotoelektrische Einrichtung schaltet das Gerät bei Erreichen des Bandendes automatisch ab, alle Funktionen können dann aber durch eine Spezial-Drucktaste "Übersteuerung" in Betrieb gebracht werden.

Geschwindigkeiten werden sowohl bei Aufzeichnung wie auch Wiedergabe durch Betätigung von Drucktasten im Bedienfeld automatisch über elektronische Festkörperschaltungen umgeschaltet. Das erübrigt zeitverschwendendes Nacheichen und Nachjustieren.

Einrichtungen für Direkt- sowie FM-Aufzeichnung und -Wiedergabe sind für alle Kanäle vorhanden, und der Frequenzgang wurde bei 152 cm/s für Direktbetrieb auf 300 kHz, für FM-Betrieb auf 20 kHz ausgedehnt.

Gleichlaufschwankungen sind über eine Bandbreite von 0,2 Hz ... 20 kHz bei 152 cm/s niedriger als 0,6 Prozent Spitze-zu-Spitze; Kompensation für die Gleichlaufschwankungen ist gegen Aufpreis lieferbar.

Die höchstmögliche mechanische Stabilität wird durch Montage der Magnetköpfe, Bandführungen und Antriebswelle auf einer Einzelplatte aus entspanntem, 38 mm dickem Aluminium gewährleistet. Die zu einer Toleranz von 0,02 mm geschliffene und polierte Oberfläche der Platte passt genau mit den exakt geschliffenen Aufbauoberflächen der Magnetköpfe zusammen. Alle Spurbreiten, Abstände, Numerierungen und Toleranzen entsprechen den IRIG-Normen.

Der PI-2100 kann in Temperaturen von +4,4°C ... +49°C bei bis zu 95% relativer Feuchtigkeit (nicht kondensierend) betrieben werden. Die Ausrüstung ist stoss- und schwingungsdämpfend aufgebaut und genügt allen militärischen Pflichtenblättern.

EE 80 753 *für weitere Einzelheiten*

Umlaufender Abtastschalter

I.D.M. Electronics Ltd, 1 Arkwright Road, Reading, Berkshire

Dieser schnellumlaufende Abtastschalter wurde von I.D.M. Electronics Ltd entwickelt, um den vielen Anforderungen in ferngesteuerten Systemen zu entsprechen. Er kann zur Lösung der normalerweise mit Telemetriearbeiten verbundenen Aufgaben eingesetzt werden, z.B. für Zerhacken von Gleichstromsignalen, Wellenformerzeugung, Verstärkerstabilisierung, Abtasten von Dehnstreifen- und Thermoelementsignalen bei niedrigem Pegel, in Digitalzählern und anderen Anwendungsmöglichkeiten, bei denen Information schnell oder langsam abgetastet wird.

Die Kombination der Edelmetall-Kontaktwerkstoffe, Schleifermaterialien und Isolierstoffe wurde ausgearbeitet, um mit sehr niedrigen elektrischen Störungen, niedriger thermischer EMK, niedrigem Antriebsmoment, hoher Isolation und langer Lebensdauer optimale Leistungsfähigkeit zu erzielen. Die mechanische Konstruktion sichert durch vorgespannte Kugellaufbahnen einwandfreies Arbeiten während der ganzen Lebensdauer des Schalters. Auf Grund seiner Konstruktion ist dieser Schalter einer der kleinsten seiner Art; besondere Aufmerksamkeit wurde der Ausdehnung seiner Lebensdauer ohne Wartung, ohne Verwendung eines Ölbades usw. gewidmet.

Die Impulslagegenauigkeit ist so hoch, dass der Schalter mit automatischen Demultiplexeinrichtungen zu verwenden ist. In HF-Systemen, in denen Koppeln zwischen Kanälen schwer auszumerzen ist, gibt dieser Schaltertyp die geringstmögliche Kopplung zwischen Segmenten.

Die Konstruktion ermöglicht eine grosse Auswahl von Anordnungen, solange das Produkt der Anzahl von Polen und Wegen 24 ist, z.B. 1polig 24wegig, 2polig 12wegig usw. Der Schalter kann durch einen Gleich- oder Wechselstrom-Miniaturmotor getrieben werden, da der Schaltersatz ein Antriebsmoment von nur wenigen Grammzentimeter benötigt und die Leistungsaufnahme selbst bei hoher Geschwindigkeit sehr niedrig ist.

Segmente kann man zur Erzielung der verschiedensten Tastverhältnisse zusammenschalten; auf Wunsch können auch spezielle Bahngestaltungen in Betracht gezogen werden. Anschlussdrähte werden entweder blank oder mit Farbkennzeichnung oder abgeschirmt, in jedem Fall aber frei zum Anschluss an die Klemmleiste des Kunden geliefert.

EE 80 754 *für weitere Einzelheiten*

Schallpegelmesser

The M.E.L. Equipment Co. Ltd, 207 Kings Cross Road, London, W.C.1

(*Abbildung Seite 269*)

Der Philips-Schallpegelmesser PM6400 wurde von The M.E.L. Equipment Co Ltd als Grundmessgerät für Ingenieure, Architekten, Techniker in der Medizin und andere eingeführt, die sich mit akustischen Messungen und Lärmbekämpfung befassen.

Das Gerät ist volltransistorisiert, daher sehr kompakt und wiegt unter 2,7 kg. Seine Tragbarkeit sowie Verwendung interner Batterien als Stromquellen erleichtern das Messen sowohl im Labor wie im Aussendienst wesentlich.

Ein Allrichtungs-Tauchspulmikrofon ist in einem Ende des Gerätes untergebracht; der Schallpegel wird auf einem Drehspulinstrument mit Skala von −7... +10 dB angezeigt. Buchsen für Anschluss eines Schreibers oder eines Filters sind vorhanden.

Das Gerät entspricht den I.E.C.-Normen und überstreicht den Bereich 33...130 dB. Bei 1 kHz ist die Unsicherheit mit dem Gerät rechtwinklig zur Schallquelle plus oder minus 1 dB. Der Frequenzbereich ist 25 Hz...15 kHz, und I.E.C.-Standardnetzwerke A, B und C sind eingebaut.

Die Stromversorgung besteht aus zwei Nickel-Kadmium-Zellen, die 20 Betriebsstunden vor Wiederaufladung gewährleisten. Ein Ladegerät PM9000 ist ebenfalls lieferbar.

Die Abmessungen sind 343 × 114 × 82.5 mm. Eine Leder-Bereitschaftstasche wird mitgeliefert.

EE 80 755 *für weitere Einzelheiten*

HF-Messender

K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(*Abbildung Seite 269*)

Der G. 30 ist ein sehr bequemer und preisgünstiger Messender, der die Frequenzen von 120 kHz...240 MHz in sieben umschaltbaren Bändern üderdeckt. Wegen seines grossen Frequenzbereiches ist er für die meisten Versuchs- und Abgleicharbeiten sehr nützlich. Die HF-Ausgangsspannung liegt über 100 mV und kann entweder intern mit 400 Hz oder extern moduliert werden. Der Modulationsgrad lässt sich mittels eines Knopfes zwischen 0...50 Prozent einstellen. Die Modulationsfrequenz von 400 Hz kann auch zwei Buchsen auf der Frontplatte entnommen werden.

EE 80 756 *für weitere Einzelheiten*

Thyristor-Steuergerät

Standard Telephones & Cables Ltd. Rectifier Equipment Division, Edinburgh Way, Harlow, Essex

(*Abbildung Seite 270*)

Ein- und Dreiphasensteuergeräte für Thyristoren, die einen Steuerimpuls mit regelbarer Verzögerung an die Steuerelektrode von Thyristoren in Stromversorgungsreglern, Drehzahlreglern für Gleichstrommotoren und ähnliche Verwendungszwecke abgeben, in denen Spannungs- oder Stromregelung erforderlich ist, werden nunmehr von STC angeboten. Die Geräte sind reichlich bemessen und auf lange Lebensdauer mit störungsfreiem Betrieb ausgelegt.

Die STC-Thyristor-Steuergeräte sind vielseitige elektronische Baugruppen, die als kompakte Bausteine für bündigen Einbau in das Bedienfeld der Silizium-Gleichrichterausrüstung geliefert werden. Eine alternative Ausführung erlaubt Montage der Steuergeräte auf Chassis innerhalb der Schränke. Verbindungen werden mittels Stecker und Buchsen, oder auf Wunsch über Klemmleisten hergestellt.

Hauptmerkmale dieser Bausteine sind:

(1) Geschlossene Stromversorgung—braucht nur ans Wechselstromnetz angeschlossen zu werden.

(2) Schnelle Ansprechzeit

(3) Rechteckwellen-Steuersignale geben vollkommene Steuerung induktiver Schaltungen.

(4) Die Impulshöhe wird durch Netzschwankungen zwischen +20 % und −50 % nicht beeinflusst.

(5) Steuert Thyristoren bis zu 2 Prozent der Höchstausgangsspannung herunter.

(6) Kann mit allen Standard-Gleichrichterverbindungen benutzt werden.

EE 80 757 *für weitere Einzelheiten*

Schutzgasschalter

B & R Relays Ltd, Temple Fields, Harlow, Essex

(*Abbildung Seite 270*)

In Erweiterung ihres Lieferprogrammes kann B. & R. Relays Ltd jetzt den Gordos-Schutzgasschalter MR201, einen mechanisch vorgespannten Umschalter, liefern. Bei einer Gesamtlänge von 78,5 m ist der Glaskörper 30,5 mm lang und hat 5,1 mm Durchmesser. Die Kontakte sind bei einem Höchststrom von 0,25 A für 10 VA bemessen. Im MR201 wird ein neuer Kontaktwerkstoff "Di-met" benutzt. "Di-met" ist ein gesetzlich geschützter Name und gibt plattierte Kontakte aus ungleichen Werkstoffen. Es ist eine Kombination von Gold für Stabilität, Kupfer für niedrigen Widerstand und Legierungszusätzen für Härte und Dauerhaftigkeit. Sie verhindern "Kleben" und "Kaltschweissen". "Di-met" hat äusserst niedrigen Kontaktwiderstand und eine nachweisbare Kapazität für Überstromspitzen beim Schalten von Wolframlampen und ähnlichen Verbrauchern. Es ist eigenstabiler als z.B. Zwischenkontaktmaterialien wie Rhodium und Silber. Die Stärke der Plattierschicht wird so dünn wie möglich gehalten, berücksichtigt jedoch Abfallen und Schaltdifferenzen. Optisch sieht die eine Zunge wie Silber, die andere wie Kupfergold aus.

Das ganze Sortiment der Gordos-Schutzgasschalter ist nunmehr entweder mit "Di-met"-Kontakten oder diffundierten Goldkontakten lieferbar.

Ausser anderen Anwendungsmöglichkeiten kann das Modell MR201 auch in die B & R-Schutzgasrelais R03 und R04 eingebaut werden.

EE 80 758 *für weitere Einzelheiten*

Miniaturübertrager

Parmeko Ltd, Percy Road, Aylestone Park, Leicester

(*Abbildung Seite 270*)

Parmeko Ltd stellt eine breite Auswahl von Miniaturübertragern und -drosseln her, die Impuls-, NF-, Netz- und Instrumentenübertrager sowie Spulen und Drosseln mit engen Toleranzen umfasst.

Den mit Baureihe 4 000 bezeichneten Typen erteilte der Ausschuss für elektronische Bauelemente der britischen Streitkräfte das Zertifikat für höhere Qualität in der Feuchtigkeitsklasse H2 und nach Temperaturkategorie 40/70 (RCS 241). Die Konstruktion dieser Übertrager und Drosseln entspricht dem Pflichtenblatt DEF 5214.

Das ganze Parmeko-Programm von Einzel- und Doppelspulenkörpern wird für die offenen Modelle der Baureihe 4 000 benutzt, die vor allem überall dort Anwendungsmöglichkeiten haben, wo es auf kompakte Zuverlässigkeit ankommt. Die meisten dieser offenen Modelle sind mit Ferritkernen lieferbar. Angeschlossen wird an in die Spulenkörperflansche gepresste Stifte. Für die verschiedenen Montagemöglichkeiten sind drei verschiedene Einfassrahmen zu haben.

Drei verschiedene Konstruktionen sind lieferbar: eingekapselt, in Aluminiumgehäuse und magnetisch geschirmt. Die eingekapselten Modelle sind in Gussharz gekapselt und haben die zusätzlichen Vorteile grösserer Starrheit und elektrischer Isolation. Die Modelle mit Aluminiumgehäuse sind mit Epoxydharz oder ähnlichen geeigneten Massen vergossen und sind—bei allen Vorteilen der eingekapselten Typen—einfacher zu montieren. Magnetische Störungen werden durch MU-Metallgehäuse abgeschirmt, und Modelle mit Zweifach- oder Dreifachabschirmung sind auf Wunsch lieferbar. Diese alternativen Ausführungen sind je nach Bauart und Anzahl der Wicklungsenden entweder mit Stiftanschlüssen oder freien Enden zu haben.

EE 80 759 *für weitere Einzelheiten*

Verstärker für Druckschaltungsmotoren

Pye Printed Motors Ltd, St. Andrews Road, Cambridge

(*Abbildung Seite 270*)

Der Motor in gedruckter Schaltungstechnik wirft durch seine Konstruktion für den Entwurf geeigneter Servoverstärker Probleme auf. Er ist im wesentlichen ein Apparat für hohen Strom und niedrige Spannung; zur Verwirklichung der vollen Konstruktionsvorteile muss ein angepasster Servoverstärker entworfen werden.

Es wird jetzt eine Verstärkerkette angeboten, die aus einem Endverstärker mit zugehörigem Treiber und Funktionsverstärker besteht und unter allen normalen Bedingungen Motoren der Baureihe 4 steuern kann, bei Begrenzung der Höchstgeschwindigkeit wegen ihrer höheren Gegen-EMK aber auch grössere Motoren.

Technische Kurzdaten für einen Endverstärker P.A.1G und den zugehörigen Treiberverstärker D.A.1G sind:

Stromversorgung
30 V− mit Mittelabgriff (±15 V−)

Dauerstrom
8,75 A_{max}

Spannungsverstärkung
D.A.1G + P.A.1G — 2x

Ausgangsnennwiderstand
0,05 Ω

Max. Verlustleistung bei 25°C Umgebungstemperatur
34 W

Für Verwendungszwecke, die höhere Ströme als 8,75 A erfordern, können zwei oder drei Endverstärkerstufen parallelgeschaltet und bis zu drei derselben durch einen Verstärker D.A.1G getrieben werden.

Der Funktsionsverstärker O.A.1G wird üblicherweise mit Gegenkopplung, die für Bestimmung der Phasenverstärkungseigenschaften der Servoschleife gewählt wird, verwendet. Der Verstärker hat folgende typische Daten:

Stromversorgung
24 V mit Mittelabgriff (±12 V−)

Ruhestromentnahme
4,5 mA

Nominelle Steuerungsverstärkung
60 — 80 dB

Stabil mit 100% Gegenkopplung

Steuerungsbandbreite
200 Hz

Drift der Eingangsspannung bei 1 kΩ Quellenwiderstand
500 µV/°C bei 25°C

Die Treiber und Funktionsverstärker sind modulare Blöcke 38×25,4×19 mm hoch und für Montage mittels Kontaktstiften auf gedruckte Schaltungen mit Rastergrundmass 2,54 mm geeignet.

Der Endverstärker ist um einen stranggepressten Kühlkörper mit Rippen aufgebaut und hat 152 × 108 × 76 mm Abmessungen.

EE 80 760 *für weitere Einzelheiten*

Induktive Tastköpfe

Electronic Machine Control Ltd, Sherman Road, Bromley North, Kent

(*Abbildung Seite 271*)

Electronic Machine Control Ltd hat ihre bestehenden Baureihen fotoelektrischer, Überschall- und kapazitiver Tastköpfe durch eine Baureihe induktiver Fühler für industrielle elektronische Regel- und Zählsysteme ergänzt. Die neuen Elemente sind in breiter Standardauswahl lieferbar und sind nichtberührende Tastköpfe, die durch Gegenwart von eisenhaltigen oder NE-Metallgegenständen beeinflusst werden. Sie sind robust gebaut, völlig abgedichtet und können unter Wasser oder in Ölen und nichtkorrodierenden Flüssigkeiten betrieben werden.

Durch ihre einfache Konstruktion können die Fühler, die keine beweglichen Teile haben, für lange Zeit ohne Wartung unter den schwierigsten Temperatur- und Schwingungsbedingungen arbeiten.

Die Abmessungen eines typischen induktiven Tastkopfes sind nur 44,5 × 22,2 mm Durchmesser. Wegen ihrer Kompaktheit eignen sie sich zum Einbau in beengte Räume, in denen andere Tastmethoden zu Schwierigkeiten oder unnötigen Komplikationen führen würden. Alle metallischen Massen, ob klein oder gross, können selbst durch Glas oder Kunststoff hindurch entdeckt werden.

Mit jedem Tastkopf wird ein elektronisches Kontrollgerät mit Netzanschluss und Abmessungen von 121 × 190 × 51 mm geliefert, das man 15 m oder sogar weiter entfernt installieren kann. Zur Erzielung langfristiger Zuverlässigkeit ist es volltransistorisiert, braucht keine Anheizzeit und—wie der Tastkops selbst—keine Wartung. Zum erstmaligen Einstellen der Empfindlichkeit ist auf der Frontabdeckung ein Empfindlichkeitsregler sowie eine Signallampe angebracht; nach der ersten Einstellung ist keine weitere Nachregelung erforderlich.

Das elektronische Gerät steuert ein internes Relais, das mit zwei für 250 V, 5 A induktionsfrei bemessenen Kontaktpaaren bestückt ist.

EE 80 761 *für weitere Einzelheiten*

Anschlussstifte

Vero Electronics Ltd, South Mill Road, Southampton, Hampshire

(Abbildung Seite 271)

Vero Electronics Ltd hat einen neuen Anschlussstift Nr. 2141 eingeführt, der mit jedem Veroboard (mit oder ohne Kupferleiter) mit Lochdurchmesser von 1,3 mm oder jeder Leiterplatte oder Anschlussleiste, die mit Löchern gleichen Durchmessers versehen sind, Verwendung finden kann. Der Kopf und die darunter angeordnete Riffelverzahnung geben einen äusserst festen Sitz, und es besteht keine Gefahr, dass der Stift beim oder nach dem Löten aus der Platte herausfällt. Diese neuen Stifte werden aus Messing gefertigt und das Löten durch Tauchverzinnen erleichtert.

EE 80 762 *für weitere Einzelheiten*

Steckeinheit mit Schalttafel

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Abbildung Seite 271)

Keyswitch Relays Ltd kann nunmehr eine preiswerte Steckeinheit mit Bauelemente-Schaltplatte Typ 304 anbieten, die Schaltungsingenieuren Benutzung eines gemeinsamen Moduls für Relais, miniaturisierte und Festkörperschaltungen, kleine Motoren und Stromversorgungen gestattet.

Die neue Einheit besteht aus dem gepressten 22poligen Sockel des Keyswitch-Relais P33, in dessen durchsichtiger Staubhaube eine Schaltplatte für Bauelemente angeordnet ist.

Um die erreichbaren Anwendungsmöglichkeiten unter Beweis zu stellen, baute Keyswitch während der Erprobungstests erfolgreich eine Ringzählerstufe komplett mit Stromversorgung, die 11 Triggerröhren, 11 Dioden, 65 Widerstände und 35 Kondensatoren umfasste, in eine Standard-Steckeinheit mit 79,4 × 56,3 × 28,1 mm Aufbauraum. Typ 304 ist auch ohne die Bauelemente-Schaltplatte erhältlich.

Wichtige Kennzeichen sind Erleichterung der Wartung und Abwandlung, da durch Ausfall von Bauelementen oder Schaltungen beschädigte Einheiten sofort vom Ersatzteillager zu ersetzen sind. Während der ursprünglichen Produktion kann man in bestehende Geräte Extrafassungen einbauen, so dass spätere Modifikationen leicht und wirtschaftlich ausgeführt werden können.

Die Einbaufassungen nehmen nur die Steckeinheiten auf, für die sie bestimmt sind, da sie durch Nasen unvertauschbar sind. Federstahl-Halteklammern in einer Auswahl von Farben gestatten Kodierung und Kennzeichnung.

Die zusammengesteckten Sockel und Fassungen sind bei einer Höchstbetriebsspannung von 750 V$\simeq$ für 10 A bemessen und haben einen Kontaktwiderstand von 5 mΩ. Der Isolationswiderstand ist bei 500 V— besser als 1 GΩ.

Die Bauelemente-Schaltplatte besteht aus Hartpapierschichtstoff mit aufgekitteten 0,04 mm starken Längskupferleitern in 3,8 oder 5,1 mm Abstand und ist mit einer regelmässigen Lochmatrize versehen. Die Verdrahtungsfläche ist 83 × 56,3 mm, und die Kupferleiter sind für 1 A mit Temperaturanstieg von 1,0°C bis zu 5 A mit Temperaturanstieg von 39,0°C bemessen. Der Widerstand ist ca. 0,018 Ω/cm und die Isolation besser als 100 MΩ. Die Höchstbetriebsspannung zwischen benachbarten Leitern ist 500 V—.

EE 80 763 *für weitere Einzelheiten*

Federleisten

Ferranti Ltd, Kings Cross Road, Dundee, Schottland

(Abbildung Seite 271)

Die Federleisten für gedruckte Schaltungen haben mit allen von Ferranti Ltd hergestellten Steckverbindungen hohe Zuverlässigkeit gemeinsam, die durch die sehr geringe Federsteife und nichtporöse Vergoldung der Kontaktflächen erreicht wird.

Die Steckfederleisten haben individuelle Kontakte geringer Federsteife, so dass die Kontaktkraft während der Fertigung in engen Grenzen kontrollierbar ist. Klammer und Feder bestehen aus Phosphorbronze, und die Klammer hat in der Berührungsfläche eine Hartgoldauflage von mindestens 0,005 mm. Die mechanische Konstruktion gewährleistet, dass unter allen Umständen mit allen vorgegebenen Plattenstärken mit der korrekten Kraft Kontakt gemacht wird. Kontaktklammern werden aus gewalztem Drahtband geformt, um scharfe Kanten zu vermeiden, die die Plattenleiter beschädigen könnten. Um ganz sicher zu gehen, hat man die Kontaktflächen gerundet.

Grosse Aufmerksamkeit wird der Kontaktvergoldung gewidmet, die laufend einem sehr hohen Standard entspricht. Muster jeder Partie werden nach DEF 5325 auf Plattierung getestet sowie Biege- und Oxydationsprüfungen unterworfen. Ausserdem werden zum Messen der Plattierungsstärke der Berührungsfläche Schnitte erstellt. In allen wichtigen Fertigungsstufen bedient man sich der statistischen Qualitätskontrolle.

Die gepressten Leistenkörper können entweder in glasgefülltem, flammensicheren Diallylphthalat oder hochwertigem, nylongefüllten Phenolharz geliefert werden. Der erstere Werkstoff gibt hohe Masshaltigkeit bei erhöhter Temperatur. Überall dort, wo die Leiterplatte mit dem Stecker in Berührung kommt, werden scharfe Radien vermieden, und die Einführung ist gefast. Polarisationsnasen lassen sich ohne Entfernung oder Beschädigung des Kontaktes in jede Polposition einsetzen.

Diese Steckverbindungen sind für 8, 16, 24, 32 und 40 Kontakte für ein- oder doppelseitige Leiterplatten, aber alle mit einer Lötfahne lieferbar. Der Nennstrom je Kontakt ist 1 A bei 100° C und die Betriebsgleichspannung 450 V. Bei Benutzung von Leiterplatten ohne Fase ist die Einsteckkraft je Kontakt ca. 156 g, bei einer Leiterplatte mit Fase jedoch nur 100 g; die gesamte Einsteckkraft ist der Anzahl Kontakte direkt proportional.

Der geregelte Kontaktdruck ist ausreichend, um den Oxydationsfilm zu durchbrechen, so dass selbst bei sehr niedrigen Betriebsspannungen der Kontaktwiderstand niedriger als 3 mΩ ist.

EE 80 764 *für weitere Einzelheiten*

Temperaturmesser

Startronic Ltd, 117a Malden Road, New Malden, Surrey

(Abbildung Seite 272)

Dies ist ein kompaktes, tragbares, batteriegespeistes Gerät zum Messen von Temperaturen von −20° C ... +120° C in zwei Bereichen von −20° ... +55° und +50° ... +120° mit ±1° C Unsicherheit.

Die als Temperaturfühler Verwendung findenden Thermistoren sind austauschbar.

Das Grossmessgerät hat ausgezeichnete Linearität und eine Skalenlänge von 95 mm.

Man kann das Gerät in beträchtlicher Entfernung vom Messkopf aufstellen, ohne die Genauigkeit zu beeinflussen. So verursacht z.B. eine Zuleitung mit 2 Ω Widerstand bei 100° C nur einen Fehler von ½° C.

Speziell angepasste Thermistoren sind lieferbar, die dem Kunden gestatten, sie —wo erforderlich—in Ausrüstungen zum Messen der Oberflächentemperatur einzubauen; auch Sonden zum Messen von Gas- und Flüssigkeitstemperaturen können geliefert werden.

Ausserdem gibt es noch zehnstufige Wählerkästchen und verschiedene Verlängerungskabel.

EE 80 765 *für weitere Einzelheiten*

Zählgerät

J.A.C. Electronics Ltd, Station Estate, Blackwater, Camberley, Surrey

(Abbildung Seite 272)

Bei J.A.C. Electronics Ltd ist jetzt die Produktion eines neuen transistorisierten Digital-Frequenzmessers und -Zeitintervallgebers angelaufen, der speziell mit einer vollregelbaren quarzgeeichten Zeitbasis entwickelt wurde, um Direktanzeige in vorgewählten Einheiten, z.B. UPM, zu geben. Zu seinen Merkmalen gehören ein hochohmiger Eingang und eine leicht lesbare einzeilige Anzeige mit einem Überschussanzeiger, so dass bei für eine vierstellige Anzeige bemessenem Aufwand eine effektive siebenstellige gegeben wird.

Zum Messen von Frequenzen zwischen 0,1 Hz und 300 kHz stehen Zählperioden zwischen 1 ms und 10 s in Millisekunden zur Verfügung; die Zählperioden können auch manuell oder durch externe Signale und/oder Schliessen von Kontakten für die allgemeine Bestimmung der Anzahl von Vorgängen in einer Zeiteinheit gesteuert werden.

Für tachometrische Aufgaben gibt es magnetische oder fotoelektrische Tastköpfe, letztere mit integralem Lampenbaustein.

EE 80 766 *für weitere Einzelheiten*

Zeitsteuergerät

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Abbildung Seite 272)

Als Teil ihres Fertigungsprogrammes für Festkörperregler hat Kent Precision Electronics Ltd ein Festkörper-Zeitsteuergerät für industrielle Anwendungsgebiete herausgebracht. Die Instrumente sind für genaue elektronische Zeitsteuerung in der Verfahrenstechnik im Bereich von 10 ms bis zu 15 s mit ±1 Prozent Gesamtunsicherheit geeignet.

Einzel- und Doppelinstrumente, die elektromagnetische Standardrelais betätigen, werden—wie in der Abbildung gezeigt — in Standard - Einbaugehäusen untergebracht. Auch Instrumente mit Impulsausgang für die Steuerung von Thyristorschaltungen für die Energieregelung bis zu Höchstnennwerten greifbarer Thyristoren sind lieferbar.

Die Instrumente beruhen auf einer konstanten RC-Zeitschaltung und einem genauen, transistorisierten Spannungsschwellenwert-Detektor, der, wenn die Spannung in der Zeitschaltung einen bestimmten Wert erreicht, schaltet und den entsprechenden Thyristorausgang zündet. Im Relaisbetrieb wird zur Steuerung eines Standardrelais ein Thyristor niedrigerer Leistung eingesetzt. Bei Festkörperregelung wird der Ausgangsimpuls für die Zündung des entsprechenden Thyristorstapels benutzt.

Eine der ansprechendsten Eigenschaften dieser Serie ist die Möglichkeit, im Standard-Einbaugehäuse ein Doppelinstrument unterzubringen. Dieses Konstruktionsmerkmal erlaubt, das Instrument so zu verschalten, dass nach Ablauf der ersten Zeitperiode die zweite eingeleitet wird. Auf diese Weise kann man das ganze Verfahren mit einstellbaren Ein-Ausverhältnissen laufend wiederholen. Andererseits kann das Doppelinstrument als zwei getrennte Zeitgeber mit konzentrischen Reglerknöpfen eingesetzt werden.

EE 80 767 *für weitere Einzelheiten*

Frequenzselektiver Verstärker

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Abbildung Seite 273)

H. Tinsley & Co Ltd stellte vor kurzem einen neuen portablen, transistorisierten, frequenzselektiven Verstärker vor, der als batteriebetriebenes Gerät viele Anwendungsmöglichkeiten hat. Er kann z.B. vorteilhaft mit Messbrücken oder induktiven Spannungsteilern eingesetzt werden, wenn es sich um Messungen bei einer Anzahl von Frequenzen handelt. Die Abmessungen sind 305 × 216 mm, das Gewicht etwa 2,3 kg.

Das Gerät kann Eingangssignale von 0,1 μV...10 V demodulieren und ist in vier Bereichen von 10 Hz...100 kHz durchstimmbar. Die Frequenz wird auf drei Dekadenskalen bis zu 10 kHz mit einer Genauigkeit von 1 Prozent der Frequenz eingestellt; der Q-Faktor ist höher als 30 und die Empfindlichkeit so gross, dass 0,1 μV über den Gesamtbereich nachgewiesen werden kann.

Drei mit geräuscharmen Transistoren mit hohem Verstärkungsfaktor bestückte Verstärkerstufen sind vorhanden. Die Spannungsverstärkung dieses Verstärkers ist in der Grössenordnung von 10^6 und ist nicht temperaturabhängig, da Siliziumtransistoren mit niedrigem Reststrom verwendet werden. Die Eingangsimpedanz des Gerätes schwankt mit der Abschwächung. Der Abschwächer ist in elf Stufen von 10...12 dB unterteilt. Für Eingangssignale über 10 μV ist die Eingangsimpedanz bei 10...50 dB Abschwächung 10 kΩ und steigt bei 50...120 dB Abschwächung für grössere Eingangssignale auf 1 MΩ.

Ein in Winkelfrequenz geeichtes alternatives Modell 5711 ist auch lieferbar.

EE 80 768 *für weitere Einzelheiten*

Tragbarer Dickenmesser

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 273)

Das jüngste Überschall-Messgerät, das von Dawe Instruments Ltd bezogen werden kann, ist der Typ 1805A Sonoray 30. Es wurde in erster Linie für Batteriebetrieb im Aussendienst entwickelt, hat aber die Genauigkeit und das Auflösungsvermögen der meisten grossen Messanlagen an Fertigungsstrassen.

Bisher wurde Überschallmessen durch den erforderlichen Netzanschluss begrenzt, der durchaus nicht immer vorhanden war und selbst dann nicht bequem über Schleppkabel zugeführt werden konnte, die oft über Leitern, Gerüste, Schächte und ähnliche Hindernisse geleitet werden mussten. Zur Lösung dieses Problemes wurde das Modell 1805A entwickelt, ein kompaktes (381 × 241 × 114 mm hoch), leichtes (7,25 kg komplett), tragbares Gerät, das normalerweise aus seinen eigenen, aufladbaren Zellen gespeist wird. Es lässt sich jedoch auch ans Netz anschliessen und ist für Rissprüfungen und Dickenmessungen geeignet.

Der Generator erzeugt Impulse mit Frequenzen von 1,5...5 MHz, deren Reflektion auf einem Schirm (63,5 × 63,5 mm) dargestellt wird. Die Skala kann gedehnt werden, bis die volle Skala einer Stahldicke von 12,7 mm entspricht. Die Messunsicherheit für Dicken im Bereich 2,5...15,25 mm ist ±0,13 mm. Bei der Risseprüfung ist die Genauigkeit nicht so wichtig, und die Reichweite wird auf bis zu 2,5 m erhöht.

Die Wandler müssen der benutzten Frequenz angepasst werden und gehören üblicherweise zur piezoelektrischen Hochleistungstype Z. Das Strandardgerät wird mit einem 1,80 m langen Anschlusskabel für den Wandler geliefert, man kann aber auch Kabel bis zu 61 m Länge verwenden, ohne dass Modifikationen erforderlich sind. Gerade Strahlung, Winkelstrahlung, Durchstrahlung und Eintauchtechnik können benutzt werden.

EE 80 769 *für weitere Einzelheiten*

Überschall-Reinigungsgerät

Sonics Division, Elliott-Automation Ltd, Bath Road, Beenham, Reading, Berkshire

(Abbildung Seite 273)

Der Bereich Schalltechnik der Elliott-Automation stellt ein neues Überschall-Reinigungsgerät vor, das sich besonders für das Reinigen empfindlicher Teile in Labors, der Feinmechanik sowie der Juwelen- und Uhrenindustrie eignet.

Das Gerät ist hochzuverlässig und bildet eine in sich geschlossene Ausrüstung. Der 2-1-Behälter (Innenabmessungen 152 × 114 × 127 mm) wie auch der 60-W-Generator sind in einem festen Gehäuse von 355 × 203 × 241 mm untergebracht. Eine Ein-Ausdrucktaste ist das einzige erforderliche Bedienelement, da Abstimmung durch Verwendung transistorisierter Schaltungen in Verbindung mit dem Elliott-Acoustica "Multipower"-Wandler überflüssig ist. Der höchste Reinigungswirkungsgrad wird sich immer und unabhängig vom Inhalt des Behälters oder Temperaturänderungen einstellen. Zum Entleeren des Behälters ist ein Abflusshahn vorhanden, und ein Schlauchanschluss an den Abfluss erlaubt Einsatz einer Pumpe und eines Filters für Umlauf, wo das wünschenswert erscheint.

Die Ausrüstung kann von ungelernten Kräften bedient werden, und die langfristige Konstanz der Ausgangsleistung gestattet Reinigungszeiten von Standarddauer.

EE 80 770 *für weitere Einzelheiten*

Überschall-Prüfgerät

C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5

(Abbildung Seite 273)

Dieses umkonstruierte Überschall-Prüfgerät, das Attenutector-Baumuster IV, ist jetzt lieferbar. Es wurde für zerstörungsfreies Testen von Kunststoff und Graphit konstruiert, ist jedoch auch für verschiedenste andere Verwendungszwecke geeignet, z.B. zum Auffinden von Blasen in Flüssigkeiten, Prüfen der

Verkittung zwischen Kunststoff und Kunststoff, Kunststoff und Metall, oder Metall und Metall, sowie zum Feststellen von Änderungen in der Dichte von Flüssigkeiten.

Wie aus dem Namen hervorgeht, arbeitet das Gerät nach dem Durchstrahlungsprinzip, nach dem das Verhältnis zwischen Abschwächung und empfangenem Signal so hoch wie 1 000 000 : 1 sein kann; um Schwankungen im Kontakt zwischen Prüfköpfen und Prüfling zu vermeiden, wird die Eintauchtechnik vorgeschrieben.

Standard-Prüfköpfe sind für Frequenzen von 250...650 kHz lieferbar und können für andere Frequenzen bis zu 2 MHz im Sonderauftrag gefertigt werden.

Das System ist im allgemeinen empfindlicher als Röntgenprüfung und sowohl in Kapitalaufwand wie Betriebskosten wesentlich billiger.

EE 80 771 *für weitere Einzelheiten*

Analog-Digitalwandler

Moore, Reed & Co. Ltd, Woodman Works, Durnsford Road, London, S.W.19

(Abbildung Seite 273)

Die von Moore, Reed & Company Ltd hergestellten Regeleinrichtungen wurden durch einen elektromechanischen Umsetzer ergänzt, in dem ein Schritt eines Fortschaltmotortriebes den Ausgang eines Verschlüsslers um eine Ziffer ändert.

Anwendungsmöglichkeiten bestehen u.a. in der Umsetzung zwei- oder dreiphasiger Wechselstromsignale in Digitalcodes, Binär- oder Gray-Code. Mit einem Subtraktor eingesetzt, erzeugt das Gerät von einem digitalen Eingang Analogsynchron-, Drehmelder- oder Potentiometerausgänge. Der Umsetzer kann auf die Impulsreihe eines Programmgerätes ansprechen, die durch die transistorisierte Logik in den den Verschlüssler treibenden Fortschaltmotor gespeist wird, um gleichzeitig analoge und digitale Signale für Regelung und Anzeige zu erhalten.

Durch Verwendung von Präzisionsgetrieben mit Vorrichtungen gegen toten Gang lassen sich die Ungenauigkeiten (Fortschaltmotoreingang - Drehmelderausgang) auf die Grössenordnung von 5 Bogenminuten reduzieren.

Die Aussenabmessungen des Gerätes sind ca. 127 × 38 × 70 mm, und es ist mit binärer V-Abtastung sowie Digitalausgängen in Gray- oder Gilham-Code lieferbar.

EE 80 772 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Elektronenrechner für Übertragungsfunktionen von R. W. Levinge

Zusammenfassung des Beitrages auf Seite 218-224

Die Leistungsfähigkeit eines linearen vierpoligen Netzwerkes oder Regelsystems wird unter allen Umständen durch seine Übertragungsfunktion bestimmt. Ein neu entwickeltes Gerät berechnet von direkten Messungen am Netzwerk oder System seine Übertragungsfunktion in Form ihrer komplexen Variablen p. In konventioneller Form beruhte die Analyse bisher auf dem Messen von Phase und Verstärkung oder des Einschwingverhaltens. Einsatz des Elektronenrechners für Übertragungsfunktionen (T.F.C.) macht die für die Ableitung der Übertragungsfunktion von Daten dieser Art aufzuwendende beträchtliche Arbeit überflüssig.

Unter der Voraussetzung, dass alle Parameter bekannt sind, kann die Übertragungsfunktion eines Systems berechnet werden. Der Elektronenrechner T.F.C. gestattet das Messen der tatsächlichen Übertragungsfunktion, wonach ausdrucksweiser Vergleich Prüfung der einzelnen Parameter ermöglicht. Andererseits ist es möglich, die Ausdrücke der bestimmten Übertragungsfunktion eines Netzwerkes oder Systems in den Rechner T.F.C. einzugeben. Vorwählbare Regler des Systems können den Daten entsprechend eingestellt werden, oder man kann den Rechner als Testausrüstung für die Annahme oder Verwerfung des Systems einsetzen.

Ein schneller Tunneldioden-Dekadenuntersetzer von M. N. S. Swamy

Zusammenfassung des Beitrages auf Seite 225-229

In diesem Beitrag wird die Konstruktion und Arbeitsweise eines Tunneldioden-Dekadenuntersetzers, der weite Toleranzen für Gleichstromvorspannung und Eingangsimpulsamplituden hat, eingehend beschrieben. Das Hauptuntersetzergerät kann bis zu 25 MHz arbeiten, die Eingangsschaltung dagegen kann nur 10 MHz bewältigen. Der Dekadenuntersetzer kann sowohl positive wie negative Impulse mit 0,5 V Amplitude abgeben.

Transistorisierter Integrator von J. A. Effenberger und R. Steele

Zusammenfassung des Beitrages auf Seite 230-234

Ein positive oder negative Sägezahnspannungen erzeugender Stromtorintegrator wird behandelt. Eine praktisch nur mit Transistoren bestückte Schaltung, die aus einem direktgekoppelten Verstärker in einem normalen Widerstands- und Kapazitäts-Rückkopplungsnetzwerk mit elektronischem Stromtor für das Entladen der Kapazität besteht, wird beschrieben und besprochen. Während des Rücklaufs fällt die maximale Ausgangsspannung von 28 V_B nach 50 μs durch Entladung auf 5 Prozent und dann nach weiteren 60 μs auf unter 10 mV. Die Linearität der Sägezahnausgangsspannung ist besser als 2×10^{-4}

Ein einfacher Transistor-Tester von M. M. Winn

Zusammenfassung des Beitrages auf Seite 235

Ein einfacher Transistor-Tester wird beschrieben. Der Transistor wird bei einem vom Bedienungspersonal gewählten Kollektorstrom geprüft und das Ergebnis direkt auf einem Messgerät angezeigt. Die Testbedingungen ähneln stark der Arbeitsweise eines Transistors als Schalter.

Ein Elektronenstromverstärker für ein Elektronen-Mikrosondenabtastsystem von J. P. Davey

Zusammenfassung des Beitrages auf Seite 236-239

Die den Signalabstand in einem Verkstärker für kleine Sekundärelektronenströme bestimmenden grundsätzlichen Faktoren werden besprochen, und es wird gezeigt, dass Elektronenmultiplikation vor Verstärkung unter gewissen Umständen keinen wesentlichen Vorteil bringt. Der Entwurf eines Verstärkers für Ströme bis zu 10^{-10} A herunter bei Bandbreiten von 0 . . . 10 kHz wird besprochen und die Arbeitsweise dieses Verstärkers beschrieben.

Ein mikroelektronischer Frequenzteiler mit veränderlichem Teilungsverhältnis von B. Preston

Zusammenfassung des Beitrages auf Seite 240-244

Die Arbeitsweise eines Frequenzteilersystems, in dem das Teilungsverhältnis auf jede beliebige ganze Nummer eingestellt werden kann, wird beschrieben (1). In einer ausführlichen Beschreibung der Schaltung werden vor allem Methoden herausgestellt, die für die Fertigung in mikroelektronischer Form von Bedeutung sind. Mit einer solchen Schaltung erzielte Ergebnisse und Wellenformen werden gegeben.

Ein neues, auf dem emittergekoppelten Multivibrator beruhendes Fourier-Synthese-Gerät von A. D. Bond

Zusammenfassung des Beitrages auf Seite 245-247

Das in dem Beitrag beschriebene Gerät erzeugt eine sinusförmige Festfrequenz-Grundwellenform und ihre Oberwellen bis zur zehnten. Alle Oberwellen sind kontinuierlich in Phase und Amplitude regelbar und können für die Synthese der verschiedensten Wellenformen summiert werden.

Gestein-Generatormagnetometer von P. I. Ghelfan

Zusammenfassung des Beitrages auf Seite 248-251

Der Beitrag beschreibt ein direktanzeigendes Gestein-Generatormagnetometer zum Messen schwacher Remanenz mit verbesserter Empfindlichkeit und leichterer Handhabung.

Eine elektronische Kompensation macht die nicht mit dem Prüfling zusammenhängende falsche magnetische Beeinflussung unwirksam. Es werden nur RC-Schaltungen verwendet, um ein schwieriges Abschirmproblem zu vermeiden.

Ein Hochspannungsimpulsgeber von I. M. Paz

Zusammenfassung des Beitrages auf Seite 252-255

Ein beschriebener Hochspannungsimpulsgeber liefert in eine kapazitive Last Impulse regelbarer Amplitude (500 . . . 1000 V), Dauer (10 . . . 110 µs) und Tastfrequenz (0 . . . 1000 Hz).

Gang- oder Spannungsänderungen der Vorder- und Hinterflanken der Impulse liegen unter 1 µs pro 1 kV.

Das Schalten wird direkt in der Hochspannungsleitung vorgenommen, die Lastkapazität während der Anstiegszeit durch die Schaltelemente geladen und während der Abfallzeit entladen.

Miniatur-Wasserstoffthyratrons ohne Vorspannung werden in diesem Gerät als Hauptschaltelemente verwendet.

Verbesserte Steuerschaltungen für steuerbare Siliziumgleichrichter von D. Ciscato und L. Mariani

Zusammenfassung des Beitrages auf Seite 256-258

Die rapide zunehmende Anwendung von steuerbaren Siliziumgleichrichtern in der Energieregelung und -umsetzung hat zur Konstruktion und Verwirklichung einer grossen Anzahl von Steuergeräten geführt.

In diesem Beitrag werden zwei Festkörperschaltungen dieser Art vorgestellt, von denen sich die eine für einphasige induktive Lasten, die andere für dreiphasengesteuerte Brücken eignet.

Ein transistorisierter Analogteiler von G. Gooberman und B. Jones

Zusammenfassung des Beitrages auf Seite 259-261

Die Konstruktion eines transistorisierten Analogteilers wird beschrieben. Der Teiler kann einen auf besser als 0,5 Prozent vorhersagbaren Ausgang abgeben, dessen Linearität für Spannungen zwischen 0,7 und 20 Volt besser als ± 0,2 Prozent ist. Ein Verfahren mit veränderlichem Tastverhältnis wird angewendet, in dem eine Tunneldiode das Kernstück des Impulsgenerators darstellt.