ELECTRONIC ENGINEERING

VOL. 37 No. 453

NOVEMBER 1965

Commentary

THE Economic Development Committee (EDC or, more popularly 'Little Neddy) for the electronic industry issued its first progress report in September of this year. This committee is one of the nine EDC's which were set up in 1964; others are being formed during the present year. The object of these committees, which form part of the larger National Economic Development Council, is to extend co-operation between Government, management and unions and to tackle the country's difficulties realistically on an industry by industry basis. The two main tasks of this particular EDC are stated to be to help formulate the electronic industry's part in the National Plan and to follow its progress and to help improve the industry's competitive power and efficiency.

It is clearly necessary that to succeed in its aims communication must be established between the Committee in London and the individual firms and people employed in the industry throughout the country. The recently issued progress report is intended as the first step in this process of communication and to this end copies of it have been sent to firms in the industry, to union branches whose members work in it and to other bodies and individuals connected with the industry. They, in their turn, have been invited to submit comments or suggestions.

The EDC for the electronic industry has as its Chairman Sir Edward Playfair, K.C.B., who was, until recently a leading personality in the computer and business machine industry. Of the eighteen members of the committee, seven are higher executives of commercial firms in the electronic industry who are sitting as representatives of the various trade associations, two are representatives of the ETU, three are representatives of the AEU, five come from Government or semi-Government departments, while the remaining member is a professor of finance and accounting. In most respects it appears to be a well chosen and broadly based committee. If there is a criticism it is perhaps that, in an industry which relies so heavily on high level scientific manpower, the professional engineer has no direct representation.

The EDC has decided, with some prompting from the Minister for Economic Affairs, that the main focus of its early work should be on imports and exports. Due to the fact that a large quantity of electronic equipment is included in other classes of goods it is difficult to arrive at reliable figures, but from those given in the progress report it is certain that some investigation is needed. According to the EDC the total imports of electronic and telecommunication equipment in 1964 were valued at about £70M, or nearly 1/16th of the apparent home market. During the same period exports amounted to some £120M. Further, overall imports have been growing about twice as fast as exports.

From an interim study of trade in electronics the EDC has made two general points. The first of these is con cerned with Government policy and several aspects of thi are mentioned. The industry's belief is restated 'that th Government in its role as customer should take accoun not simply of direct financial criteria but also of the wide implications of its expenditure for the industry's technica progress and international competitiveness'. In othe words the Government should give a lead in 'buyin British' and should also give encouragement to Britis manufacturers to design and develop the sophisticate equipments and products that are needed in the moder world. This, of course, is no new idea, it has been state and restated many, many times; nor is it confined to th electronic industry. It is a sentiment which, for example would be strongly echoed by the aircraft industry. Th EDC also stresses the importance of Government policie in other respects and gives as examples, education an research. In connexion with this latter point the industry' view is stated 'that research and development in the field where the Government is the main purchaser shoul as far as possible be contracted out to industry and that th research and development done within the Governmen should fit in with a common policy worked out in col laboration by Government and industry.' This of cours raises a number of difficult problems, the solution o which could be highly beneficial. To avoid completel duplication of research would be disastrous, but to attai a far higher degree of co-ordination would save a grea deal of time, money and above all valuable scientific man power. Perhaps what is really needed is a considerabl more scientific and organized approach to the dissemina tion of knowledge than exists at present.

The second general point made by the EDC is the nee for firms to think in terms of export markets and it i continuing with a study of the prospects and problem for exporters and expects to formulate action proposal by the end of the year. In these deliberations, particula emphasis is being given to the particular problems of th smaller firms and the methods by which these may b overcome.

The EDC is also studying ways in which efficiency ca be increased, how greater liaison with other industrie can be brought about and the future manpower require ments of the industry.

No radically new ideas have yet been propounded bu this could hardly be expected after so short a time. How ever, a start has been made and some fruitful avenues fo investigation have been determined. If the members of th EDC are to achieve their objectives they have no eas task ahead of them and one can but wish them well i their onerous labours.

White Noise Loading of Multi-Channel Communication Systems

By W. Oliver*

This article discusses a test method now used internationally for testing multi-channel communication systems. It shows that Gaussian noise has the same statistical properties as the multi-channel signal, and gives methods of converting results obtained from noise power ratio measurements to signal-to-noise-ratio etc.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 781)

WHEN a large number of telephone channels are to be carried by coaxial cables or by a broad band radio relay link, it is now a universal practice to use frequency division multiplexing with 4kc/s channel spacing for base band loading of the system. The number of channels provided is usually an integral multiple of twelve.

The resulting multiplex signal comprises many rapidly changing components of various frequencies and if any non-linear distortion in the transmission system exists, this will introduce unwanted intermodulation products of these components. When distortion is severe, unwanted products are heard by the telephone subscriber as a background signal resembling random noise. These spurious noise products are measured with a psophometer in or related to, a telephone test channel in picowatts (pW) with weighting applied to approximate to the relative interfering effect of various frequency components. (Note 1pW = 10^{-12}W).

Standards of transmission performance for long distance radio circuits have been suggested by C.C.I.R. (International Radio Consultative Committee), and D.C.A. (Defense Communications Agency); for long distance cable circuits by C.C.I.T.T., (The International Telegraph and Telephone Consultative Committee). For the purpose of this article, the most important of these standards are the recommendations for 2 500km (1 562·5 miles) and 6 000 nautical mile telephone circuits. The 2 500km recommendations state that the noise measured at a point of zero relative level in any audio channel shall not exceed 10 000pW psophometric for more than 1 per cent of the busy hour. Of this total, 2 500pW is allotted to the terminal multiplexing equipment and the remainder (7 500pW) is the maximum allowed for all forms of noise and intermodulation introduced in the transmission path.

The DCA 6 000 nautical mile recommendations state that the spurious noise should not exceed 25 000pW psophometric during the busiest hour in a month of poorest transmission. Of this total, 20 000pW is assigned to the transmission equipment and 5 000pW to the multiplex.

These recommendations provide an excellent method of specifying performance from an individual telephone user's point of view, but they are quite impractical for a precise specification of equipment performance, for acceptance testing of new systems, or for general routine testing. For these purposes, the spurious noise requirements must be translated into a test method which can be carried out rapidly under accurately controlled conditions with repeatable results.

The method now being internationally used is to load up the base band of the equipment with random noise to simulate traffic conditions, to introduce perhaps three quiet slots in the noise spectrum at the transmitter and to measure, at the receiving terminal, the resulting noise in the slots due to distortion and propagation effects. This is commonly called White Noise Loading.

Load Rating Theory of Multi-Channel Amplifiers

In a perfect multi-channel carrier telephone system each channel would be entirely free from interference produced by the energy present in the other channels. Since all the channels are amplified by the same repeaters, which as a practical matter cannot have perfectly linear characteristics nor be free from crosstalk effects, this ideal may be approached but never completely realized.

The inter-channel interference must be kept below a

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value which will be satisfactory for the grade of transmission concerned; further reduction is uneconomic.

Single- and Multi-Channel Amplifiers

In a single channel amplifier, the loading depends only on the channel input, the ratio of peak to r.m.s. signal is about 19dB with speech inputs. The effect of non-linearity in an amplifier carrying a single channel signal is to produce harmonic and intermodulation distortion of the channel signal only.

In a multi-channel amplifier the loading depends upon:

(a) The number of active channels.

(b) Talker volume averaged out over the number of active channels.

(c) The presence of pilot signals required to synchronize the carrier equipment.

(d) High level 'ringing' tones, used to establish initial communication.

There is a diversity factor which reduces the per channel loading as the number of channels increases. As a result, the ratio of peak to the r.m.s. value of the signal tends to 13dB when there is more than 60 active channels.

The distortion components falling into a given channel are generally unrelated to the signal in that channel; they are due to the loading on the other active channels since the distortion introduced on the channel signal itself is small by comparison. As a result, these components cannot be regarded as distortion of the channel signal, but rather as random interference or *noise*.

Multi-Channel Instantaneous Voltage Distribution

Fig. 1 shows the variation in peak to r.m.s. voltage of a multi-channel signal as the number of active channels is varied. It can be seen that when N is greater than 16, the 'normal' or Gaussian distribution curve has a very similar shape to the telephone channel curves. This is also the distribution curve of Gaussian or White Noise.

Multi-Channel Peak Factor

As discussed previously, as the number of active channels increases, the peak to r.m.s. ratio of the multi-channel signal varies from 19dB for a single channel to 13dB when there are more than 60 active channels.

It should be noted that the peak to r.m.s. ratio of Gaussian noise is also 13dB. The characteristics of multi-channel signals and Gaussian noise are such that the only way that their amplitude can be expressed is in terms of amplitude probability. This is obtained by measuring the percentage of time that the peak value exceeds the r.m.s. value by a given number of decibels. The peak factors quoted refer to measurements for 0·001 per cent of time.

TABLE 1
Activity Coefficient

NUMBER OF CHANNELS NOMINAL N	ACTIVE n	ACTIVITY COEFFICIENT k
12	7	0·58
24	11	0·47
60	23	0·38
120	41	0·34
240	77	0·32
600	180	0·30
960	288	0·30

NOTE: *The activity coefficient of systems greater than 600 channels is a constant (0·3 from CCIR), although Holbrook and Dixon[2] quote 0·25 as their measured figure.*

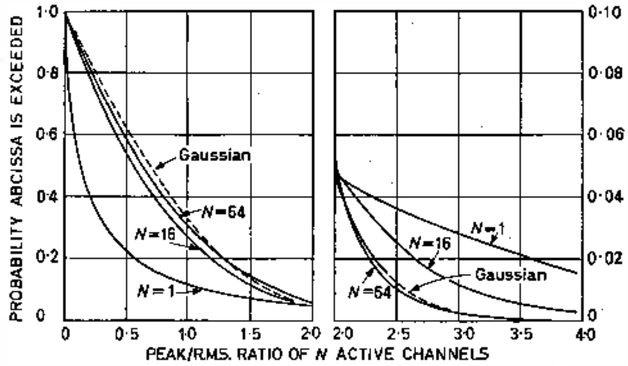


Fig. 1. Variation in peak to r.m.s. voltage

ACTIVITY COEFFICIENT WITH TELEPHONE LOADING

With normal telephone conversation loading of a multi-channel system, none of the channels has a continuous input applied because:

(a) When a subscriber is talking, his correspondent is listening.

(b) The subscriber does not talk continuously, there will be punctuation pauses.

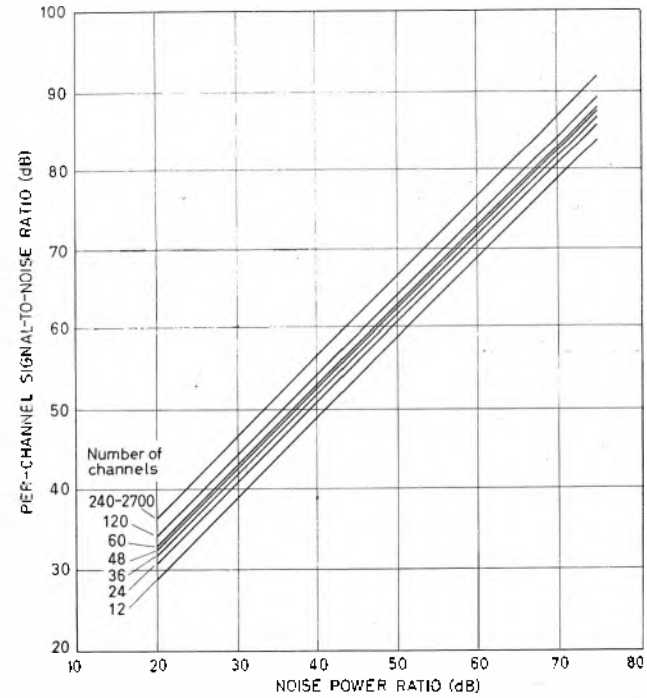
(c) A system is usually designed so that there are channels in reserve even during 'busy hour' conditions.

The activity coefficient of a system is the fraction of the system capacity that is being actively used at a given time. Values quoted were measured statistically for greater than 1 per cent of time. It will be noted from Table 1 that systems having few channels have the highest activity coefficient, this occurs because the averaging effect between channels is less.

ACTIVITY COEFFICIENT WITH DATA LOADING

With data or teleprinter loading of a multi-channel system the activity coefficient is much higher than with telephone loading. The trend is to assume 100 per cent loading at all times.

Fig. 2. N.P.R. to signal-to-noise ratio conversion



Mean Power Level of the Multi-Channel Signal

With Telephone Loading

C.C.I.R. recommend two standard equations to approximate to the complex signal mean power curves produced by Holbrook and Dixon[2]; these are:

For systems with less than 240 channels

$P_N = -1+4 \log N$ dBm0

(dBm0 = dBm referred to point of zero relative transmission level)

For systems with greater than 240 channels

$P_N = -15+10 \log N$ dBm0

where P_N = Mean power of the multi-channel signal

N = Nominal number of channels.

These equations allow for variation in activity coefficient and multi-channel peak factor, they were derived empirically from actual measurements of multi-channel systems (see Holbrook and Dixon[2]).

With Data Loading

The D.C.A. standard equation for data loading is:

$P_N = -5+10 \log N$ dBm0

This applies to all values of N.

Noise Loading Techniques

Simulating the Multi-Channel Signal with White Noise

It was shown previously that the statistical properties of Gaussian or white noise are very similar to those of the complex multi-channel signal when there are more than 16 active telephone channels. As a result, it is a legitimate measuring technique to use white noise to simulate the multi-channel signal.

The required noise power to simulate N channels can be obtained from the equations shown previously which are:

$P_N = -1+4 \log N$ dBm0 for N less than 240 channels

$P_N = -15+10 \log N$ dBm0 for N less than 240 channels

and $P_N = -5+\log N$ dBm0 for data loading

where P_N = Noise power required to simulate complex signal power due to N telephone channels.

Noise Bandwidth

To test a communication system with noise loading it is necessary to limit the bandwidth of the noise to that of the system baseband. The range of baseband frequencies that occur can be seen from the Filter Selection Chart (Table 2).

This is accomplished in Marconi noise loading sets by applying the white noise to a high-pass filter to define the lower end then to a low-pass filter to set the high frequency end of the noise spectrum.

Measurement of Intermodulation and Thermal Noise Using White Noise Loading

As mentioned previously, the effects of distortion in multi-channel amplifiers is to produce interference resembling random noise in the measured telephone channel. The method used with white noise loading to measure distortion and other noise components is to place a number of 'quiet' slots in the white noise spectrum before it is applied to the baseband of the system. The resulting 'spillover' noise into the noise slot is measured at the baseband output of the remote receiver. The noise measured will be due to distortion effects and thermal noise of the entire communication system from the baseband input of the sending station to the baseband output of the remote receiver.

If three measuring slots are used, one near the low end, one in the middle and one at the upper end of the baseband, the results obtained will show the distribution of spurious components over the entire baseband.

Noise Power Ratio

Marconi noise loading sets measure the intermodulation and thermal noise of a system in terms of n.p.r. (noise power ratio).

N.P.R. is defined as follows: Noise in a test channel with all channels loaded with white noise, to noise in the test channel with all channels except the test channel fully noise loaded, expressed in decibels.

Baseband Intrinsic Noise Ratio

The Baseband Intrinsic Noise Ratio (b.i.n.r.) of a system is a measure of idle noise due to thermal noise, carrier leaks, spurious f.m. in oscillators, etc.

B.I.N.R. is defined as noise in a test channel with all channels noise loaded, to noise in the test channel with all noise loading removed.

The difference between measured n.p.r. and b.i.n.r. indicates the amount of spurious noise in the system caused by intermodulation distortion and crosstalk.

In addition to a b.i.n.r. measurement of idle noise, it is normal practice to examine the baseband signal with a spectrum analyser or selective voltmeter to ensure that the b.i.n.r. reading is not in error due to the presence of spurious signals (e.g. excessive carrier leaks, etc.).

Conversion of N.P.R. to Single Channel Signal-to-Noise Ratio

To relate the n.p.r. reading of a noise loading test to equivalent channel signal to noise ratio or test tone level to noise, the following conditions apply:

$N = P_G \,.\, B_T/B_G \,.\, 1/P_R$

where

N = Noise power ratio (numeric)

P_G = Power output from noise generator or power applied to receiver (mW)

B_T = Effective bandwidth of a telephone channel (300 to 3 400c/s)

B_G = Bandwidth of noise generator (kc/s)

P_R = Noise power due to intermodulation and system noise in the telephone channel.

Re-writing the previous expression:

$$P_R = P_G/N \,.\, B_T/B_G$$

or

$$1/P_R = \frac{N\,B_G}{P_G\,B_T}$$

$$\therefore\ T/P_R = \frac{T.N.B_G}{P_G\,B_T} = N \,.\, B_G/B_T \,.\, \frac{1}{P_G/T}$$

TABLE 2

Typical Filter Frequencies for Various System Capacities

NO. OF CHANNELS	BAND LIMITS HIGH-PASS (kc/s)	BAND LIMITS LOW-PASS (kc/s)	IN-BAND SLOTS LOWER (kc/s)	IN-BAND SLOTS CENTRE (kc/s)	IN-BAND SLOTS UPPER (kc/s)
12	12	60	27	40	50
24	12	108	40	70	105
36	12	156	40	70	105
48	12	204	40	105	185
60	12	252	40	185	245
60	60	300	70	185	270
120	60	552	70	270	534
240	60	1052	70	534	1002
300	60	1300	70	534	1248
600	60	2660	70	1248	2438
960	60	4028	70	2438	3886
960	316	4188	534	2438	3886
1200	316	5564	534	3886	5340
1800	316	8204	534	3886	8002
2700	316	12388	534	3886	12150

where T = Test Tone Level of a telephone channel (dBm)
$10 \log (T/P_R) = 10 \log N + 10 \log (B_G/B_T) - 10 \log (P_G/T)$
or Test Tone Level to Noise Ratio in decibels
= n.p.r. (dB) + 10 log (B_G/B_T) − ratio (dB) of noise generator power to test tone level.

(The last term is commonly called the Noise Loading Ratio and is given by the C.C.I.R. loading equations $P_N = -15 + 10 \log N$, etc.)

The telephone channel whose signal-to-noise ratio is calculated by this method is the channel that the baseband slot frequency is translated to by the multiplex equipment.

In practice, the measured signal-to-noise ratio of a channel will be lower than that calculated because of additional crosstalk and intermodulation effects in the multiplex equipment. Fig. 2 shows graphically n.p.r. to signal-to-noise ratio conversion.

A modern white noise test set is shown in the photograph on page 714. It is fully transistorized and is housed in two compact units. It is designed for testing cable and radio multi-channel links with capacities up to 2 700 channels.

Conversion of Signal-to-Noise Ratio to Channel Noise

When the measured n.p.r. of a system has been converted to channel signal-to-noise ratio, the channel noise can be calculated as follows:

(a) *dBa0* Noise Units

The term dBa refers to F.I.A. weighted noise measurement above a reference noise level of −85dBm, i.e. 0dBa = −85dBm (at 1kc/s). However, when white noise is applied to an F.I.A. weighting filter it is attenuated by 3dB

0dBa = −85 + 3 = −82dBm (white noise applied)

∴ To convert signal-to-noise ratio to dBa0
dBa0 = 82 − signal-to-noise ratio.

(b) *dBrnc0* Noise Units

The term dBrnc refers to 'C' message weighted noi measurements above a reference noise level −90dBm, i.e. 0dBrnc = −90dBm (at 1kc/s). Whe white noise is applied to a 'C' message weightir filter, it is attenuated by 1·5dB.

0dBrnc = −88·5dBm (white noise applied)

∴ To convert signal-to-noise ratio to dBrnc0
dBrnc0 = 88·5 − signal-to-noise ratio.

(c) Picowatts of Noise Power

$1\text{pW} = 10^{-12}\text{W}$ or $0\text{dBm} = 10^{9}\text{pW}$

Picowatts of noise power are usually measured wi psophometric weighting. When white noise is applie to a C.C.I.R. psophometric weighting filter it attenuated by 2·5dB (almost the same as F.I. weighting).

Assuming 0dBm0 = 0dBm at the point of measur ment, to convert signal-to-noise ratio to picowatts

$$\text{pW} = \frac{10^9}{\text{Antilog signal-to-noise ratio}}$$

$$\text{or pWp} = \frac{10^9 \times 0{\cdot}56}{\text{Antilog signal-to-noise ratio}}$$

If 0dBm0 is not equal to 0dBm, the noise pow should be corrected to suit the difference in level.

REFERENCES

1. Fagot, J., Magne, P. Frequency Modulation Theory with Microwa Link Applications. (Pergamon Press, N.Y. 1962).
2. Holbrook, B. P., Dixon, J. T. Load Rating Theory for Multi-Chann Amplifiers. *Bell Syst. Tech. J. 18* (Oct. 1939).

New Communication Equipment

Among the recently introduced communication products of Racal Communications Ltd are a fully transistorized h.f. receiver type RA329 and an h.f. manpack transmitter/receiver known as 'Squadcal'.

The RA329 s.s.b./f.s.k. receiver has been developed to meet the requirements of military and quasi-military services for a compact, self-contained, transportable s.s.b. and teleprinter receiving terminal. The receiver covers the frequency range of 1 to 30Mc/s with an in-line digital scale, the 1kc/s display providing interpolation calibration of 200c/s. The receiving terminal comprises an RA.217 receiver, together with an f.s.k. unit, loudspeaker/amplifier unit and connector panel. These units are mounted in a rugged frame which, in turn, is housed in an approved type military field transit case. A sealed lid is fitted for protection against ingress of dust or moisture during transportation, and is easily detached and stowed to enable the receiving terminal to be rapidly put into operation.

The basis of the RA.329 terminal is the RA.217 high stability h.f. receiver, employing the 'Wadley' drift-cancelling system of frequency-changing and selection. Separate demodulators are incorporated for s.s.b./c.w. and d.s.b. reception, a product detector being employed with separate reinsertion oscillators for the selection of u.s.b. and l.s.b. with variable b.f.o. for c.w. reception.

Alternative i.f. bandwidths of 200c/s, 1kc/s, 3·5kc/s and 13kc/s are provided. The f.s.k. adaptaor is suitable for reception of frequency shift signals of 200c/s to 1 000c/s at speeds of up to 100 baud, and also contains an a.f. amplifier/monitor loudspeaker unit. The connector panel provides for external connexion to an l.f. convertor, panoramic adaptor, synthesizer or frequency counter etc. The power supply unit permits operation from a.c. supplies of 100V to 250V, 45 to 400c/s and from a nominal 24V d.c. supply with positive or negative earth. The RA.329 has been designed to meet Defence Specification DEF.133 for operation in ambient temperatures of +55° to −10°C.

'Squadcal' is a fully transistorized s.s.b. transmitter/receiver designed for h.f. communications for military, police and security forces and is suitable for operation under all enviro mental conditions.

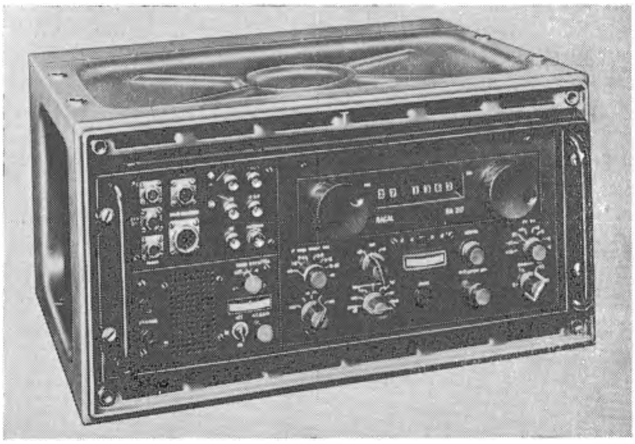
Suitable for use with dipole and sloping wire aerials, addition to an 8ft whip aerial, 'Squadcal' can be employ both for short-range ground-wave use and for long-distan sky-wave operation. 29 channels in the frequency range of 2 7Mc/s are available for s.s.b. telephony, key-toned telegrap (normally u.s.b. but l.s.b. can be provided as optional.

The transmitter provides an output of 5W p.e.p. with carrier suppression of 40dB with a power consumption 550mA at 18V.

The receiver has a sensitivity of 1µV for a 5mV aud output with a signal-to-noise ratio of 15dB.

The selectivity is −6dB at 2·7kc/s and −40dB at 5·6kc/ and the power consumption for 10mW audio output is 80m at 18V.

The RA 329 receiver



Shot Noise and the Measurement of the Electron Charge

By E. Mathieson*

Experimental details are given of a measurement of the shot noise in the current of a thermionic diode. From this measurement a value for the electron charge 1·62 . 10^{-19} coulombs was obtained, the experimental uncertainty in this result being about 5 per cent. The role of this type of investigation as a student laboratory experiment is discussed, and a simple account of the relevant theory is given.

(Voir page 774 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 718)

N experiment which deserves consideration for an advanced level physics course is a measurement of ɔt noise. The effect, first predicted and described by ottky[1], is fundamental, arising directly from the ɔmicity, or graininess, of electricity. An early acquaint-ɔe with practical limitations on electrical measurements e to noise is desirable, and in this case quantitative ɜasurements of the noise can be made relatively easily. ese measurements can not only yield a value for the ctron charge, but they also afford an instructive, mean-;ful exercise in a.c. techniques.

As long ago as 1925 in an admirably conceived and ɔcuted experiment Hull and Williams[2] examined criti-ly Shottky's theoretical predictions. Although at that ıe they had to design and construct relatively unusual paratus, such as amplifiers and attenuators, they were vertheless able to make a fairly precise measurement of electron charge and also to show the smoothing effect shot noise of space charge. Today one can expect to d suitable apparatus to perform this type of investiga-n (with varying degrees of precision) in any moderately ll equipped teaching laboratory. This article describes particular experimental arrangement, to be included in student laboratory course, for evaluating the electron ırge from measurements of the shot effect in a diode.

ıe Shot Effect

The current through a thermionic diode is of course e to a flow of electrons each having a finite and fixed ırge. Under temperature limited conditions the emission each electron is an independent event; the resultant :rent will therefore show fluctuations about the average lue according to a Poisson distribution (see, for instance, ans[3]).

Let I be the average diode current and e the magnitude the electron charge. Then $I\Delta t/e$ represents the average mber of electrons emitted in a time interval Δt. Accord-; to Poisson statistics, the root mean square fluctuation out this average number is just $(I\Delta t/e)^{\frac{1}{2}}$. This will be :ognized as the same type of statistical fluctuation as countered in most nuclear counting experiments. The ɔt mean square current fluctuation is then $(Ie/\Delta t)^{\frac{1}{2}}$, d it is important to appreciate that the value of this ı.s. noise current depends upon the time interval Δt ɔsen for current measurement. Shottky pointed out that time interval is effectively defined by passing the diode rrent through a tuned circuit. More generally, a definite ıe interval is implied by restricting the noise measure-:nt to a finite frequency bandwidth, and it can be shown e, for instance, MacDonald[4]), that in a bandwidth Δf r.m.s. shot noise current is $(2eI\Delta f)^{\frac{1}{2}}$.

'he University, Leicester.

Although this last result is obtained by important and rather general methods, in fact the special case of interest here can be treated most rapidly and easily by direct application of Campbell's theorem (see Appendix). A result of this theorem is that if the response of a system to a single, extremely short duration, impulse is $f(t)$ then

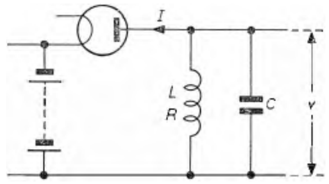


Fig. 1. Basic circuit for the calculation of shot noise voltage

the mean square deviation of its response to an average rate of N equal magnitude impulses per second arriving randomly is $\langle v^2\rangle$ given by

$$\langle v^2\rangle = N \int_0^\infty f^2(t)dt$$

Now the present system to be analysed is shown schematically in Fig. 1, where it is desired to calculate the mean square shot noise voltage $\langle v^2\rangle$ developed across the parallel LC circuit by the diode current. In this case the single impulse is the arrival at the diode anode of an individual electron of charge e and the subsequent response $f(t)$ of the system is a damped cosine wave voltage. (In practice the damping must be light, and this condition will be assumed in the following analysis.)

Thus

$$f(t) = \frac{e}{C}\exp\left(-\frac{R}{2L}t\right)\cos\left(\frac{1}{LC} - \frac{R^2}{4L^2}\right)^{\frac{1}{2}} t$$

$$= \frac{e}{C}\exp\left(-\frac{R}{2L}t\right)\cos\omega t$$

and

$$\int_0^\infty f^2(t)dt = \left(\frac{e}{C}\right)^2\frac{L}{2R}\left(1 - \frac{1}{4Q^2}\right)$$

where

$$Q = \frac{\omega L}{R} \simeq \frac{1}{\omega CR}$$

As stated above $Q \gg 1$ (in the present experiment $Q \simeq 36$), so that $1/4Q^2$ may be justifiably neglected compared with unity.

$$\text{Hence } \langle v^2\rangle = N\left(\frac{e}{C}\right)^2\frac{L}{2R} \text{ where } N = \frac{I}{e}$$

$$\text{Thus } \langle v^2\rangle = \frac{IeL}{2C^2R}$$

This last is the equation to be examined. It was first derived by Shottky, and subsequently by many others. The resistance R includes, of course, the effect of any resistances shunting the parallel circuit. If the diode is operating under a true temperature limited condition its a.c. resistance should be infinite and R should be independent of average diode current. An accurately linear relationship between $\langle v^2 \rangle$ and I has therefore been taken, in this experiment, as an indication that this condition has been obtained.

It should be noted that, unlike the Millikan experiment, this experiment cannot verify the uniqueness of the electron charge. A unique value is assumed, and experimental verification of the Shottky equation yields this value numerically. Thus if electrons were always emitted in pairs, say, then the experiment would yield a value twice the electron charge.

Description of Apparatus and the Measurement of Circuit Constants

An experiment to verify the Shottky equation or, assuming its correctness, to measure the electron charge must pay particular attention to three aspects. Firstly, the diode must be operated under temperature limited conditions. The presence of space charge rapidly reduces the shot noise below the value given above because electron transits are then no longer independent events. Secondly, the noise must be detected by a true square law detector; a valve-voltmeter for instance is unsuitable for this task. Thirdly, the amplifier used to amplify the shot noise must be operated at a sufficiently low level that overloading on unusually large noise excursions is negligible. Another important requirement, in fact easily obtained, is that the amplifier should have a flat response over the frequency range covered by the tuned circuit. The circuit arrangement employed for the present experiment is shown in Fig. 2. The tube generating the shot noise was a Ferranti coaxial diode, type GRD7, having a tungsten filament cathode of diameter 0·125mm. The potential drop across the diode was maintained at 120V by a dry battery, and the diode current was varied by varying the filament current, (~1·5A).

A cathode-follower was used to match the tuned circuit to the coaxial cable and low input impedance of the main amplifier. A compound connected emitter-follower was used at first, for greater convenience, but this was replaced because of possible drift in the value of input impedance. The overall voltage sensitivity of the system could be measured by switching the cathode-follower input to the 75Ω load of a wide band attenuator and signal generator,

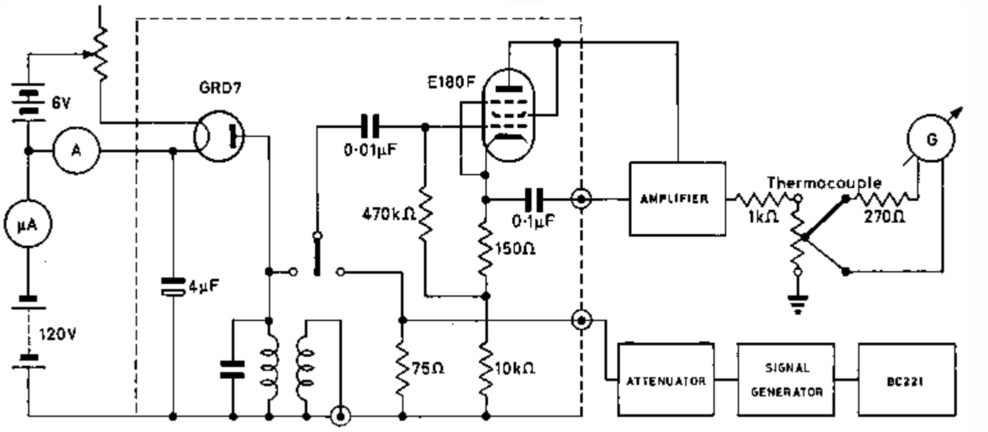


Fig. 2. Experimental arrangement used for the measurement of shot noise

as shown. The amplifier employed was a nuclear pul amplifier with maximum gain 50 000, input impedan 2kΩ, output impedance 20Ω, and maximum outp ±100V. Its response was found to be essentially flat the range 300 to 400kc/s. While taking the measuremen described in this article the amplifier gain was approx mately 16 000.

The amplifier output was fed via a 1kΩ resistor to t 32Ω heater of a vacuum thermocouple[5]. The thermocouy itself was in series with a 270Ω resistor and a 'Scalam galvanometer of nominal sensitivity 7·6cm/μA a resistance 90Ω. The sensitivity of this arrangement w such that full scale (14cm) deflexion was obtained for amplifier output of slightly less than 3·3V r.m.s. The pro ability of amplifier output over-loading was thus neg gible. Preliminary experiments were necessary to establi the range of square law response of the thermocoup These experiments, and the sensitivity of the galvanomet determined the values for the series resistors.

The diode, tuned circuit and matching circuit we enclosed in an aluminium shielding box, and coaxi cables were used to conne to the main amplifier and the step attenuator and sign generator. The batteries a filament control circuit f the diode, and the therm couple and galvanomet could however be left u shielded without any dete able instability. The amplifi and signal generator had se contained stabilized pow supplies, and their a supplies were obtained fr a stabilized source.

In contrast to the origir arrangement of Hull a Williams, the tuned circ consisted of a large inductance (2·10mH) and small capacitance (91·6pF). This capacitance w partly that of a fixed capacitor of nominal val 68pF but included also the diode capacitance, the inp capacitance to the cathode-follower, and wiring and str capacitances. The circuit resonated in situ at 363·0kc This frequency was determined by injecting a signal fr a coil of a few turns very loosely coupled to the indu ance and tuning the signal generator through the reso ance peak observed on the thermocouple galvanomet The circuit capacitance C was measured by noting t resonance frequency f_r, and the new resonance frequen f_r', obtained by connecting in parallel to the tuned circ an accurately known standard capacitor C_s (105·2pF). Th $C_s/C = (f_r/f_r')^2 - 1$. From a knowledge of f_r and C t circuit inductance L could be calculated.

The circuit resistance was measured by examining t shape of the resonance curve. Suppose a constant amp tude signal is injected into the tuned circuit from t loosely coupled coil and signal generator. Then, provid that the overall sensitivity of the system follows accurat a square law, the ratio r or the response at resonance that at an angular frequency ω is given by

$$r = 1 + \left(\frac{\omega^2 LC - 1}{\omega RC}\right)^2$$

Then $$(r-1)^{\frac{1}{2}}\omega RC = \left(\frac{\omega}{\omega_r}\right)^2 - 1 \simeq 2\frac{\omega - \omega_r}{\omega_r}$$

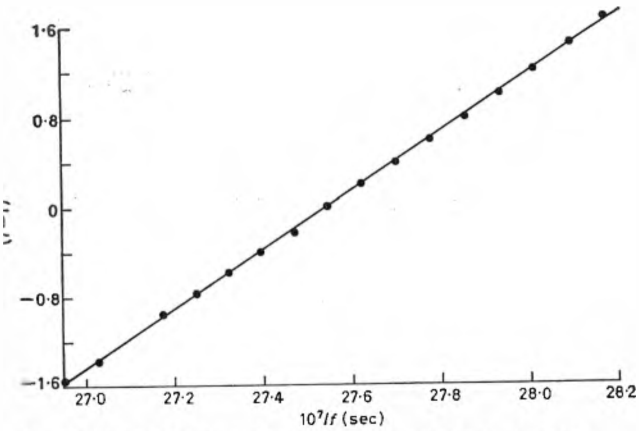


Fig. 3. Plot of experimental points used for the evaluation of ***R***

ınder the conditions of the present experiment this last pproximation represents a departure from equality of ss than 0·7 per cent for $(f - f_r) = 5$kc/s. The resultant ror in the calculation of R is much smaller than this ecause of the form of the final equation. Thus, with this pproximation

$$(r - 1)^{\frac{1}{2}} = \frac{1}{\pi RCf_r} - \frac{1}{\pi RCf}$$

he Q of the tuned circuit was about 36 (10kc/s bandidth between 3dB points). With so sharp a resonance an curate measurement of frequency was essential, and gnal generator dial readings were quite inadequate. ortunately, an army surplus crystal frequency meter, type C 221, was available so that the signal generator freiency could be measured to within about 30c/s in this equency range.

Fig. 3 shows a plot of $(r - 1)^{\frac{1}{2}}$ against $1/f$. The gradient the best straight line through these experimental points as calculated, leading to a value for the circuit resistance of 131·6Ω (a consideration of errors is given later). It interesting to note that the d.c. resistance of the coil was pproximately 42Ω.

ıot Noise Measurement and Amplifier Calibration

Under temperature limited conditions, the shot noise sults in a galvanometer deflexion θ related to the diode ırrent I by the equation

$$\theta = S \frac{IeL}{2C^2R} + \theta_0$$

here S is the overall sensitivity of the system (cm/V^2), ıd θ_0 is the deflexion due to (constant) internally nerated noise. Fig. 4 shows the result of plotting θ ainst I in the range 0 to 240μA. The accurately linear lationship referred to previously is illustrated here; the lculated gradient of the best straight line through the perimental points is $4{\cdot}99.10^4$cm/A. If the range of I is creased to 1mA (with a corresponding switched decrease amplifier gain) a very slight curvature can be discerned the plot; if the range is further increased to 5mA a arked curvature is found. The current range over which on-linearity becomes apparent depends upon the geoetry of the particular diode employed. The initial dexion θ_0 corresponds to an r.m.s. noise referred to the put of about 20μV.

The overall sensitivity S of the system was measured the following manner. With the signal generator as a urce and with the step attenuator correctly matched, a libration curve of the system was obtained in terms of e (unknown) output of the generator. The thermocouple heater was then disconnected from the amplifier output (a second thermocouple would have been preferable) and, with a 43Ω resistor in series, was connected across the attenuator output. The galvanometer deflexions for a range of attenuator settings were noted. The generator was then replaced by an accurately known and stable d.c. voltage source, (1·00V), and the attenuator adjusted for a similar range of galvanometer deflexions. Thus, in this way the measured d.c. sensitivity of the thermocouple and attenuator is used to determine the a.c. output of the signal generator. It is, of course, necessary that the generator output remains stable during this calibration procedure. For the present experiment the generator output was found to be 124mV, and the overall sensitivity of the system was calculated at $3{\cdot}26.10^8$cm/V^2. It was found that inconsistent results were obtained in this calibration procedure if poor quality coaxial plugs and sockets were employed at the 75Ω level.

By substituting the appropriate numerical values, quoted in this and in the previous section, into the shot noise equation the value found for e was $1{\cdot}62.10^{-19}$ coulombs. The experimental error in this result is discussed in the next section.

It may be helpful to summarize here the main stages of measurement in the experiment.

(1) Preliminary adjustments and experiments to establish (or verify) the square law response of the system over a suitable range.

(2) Measurement of C and L.

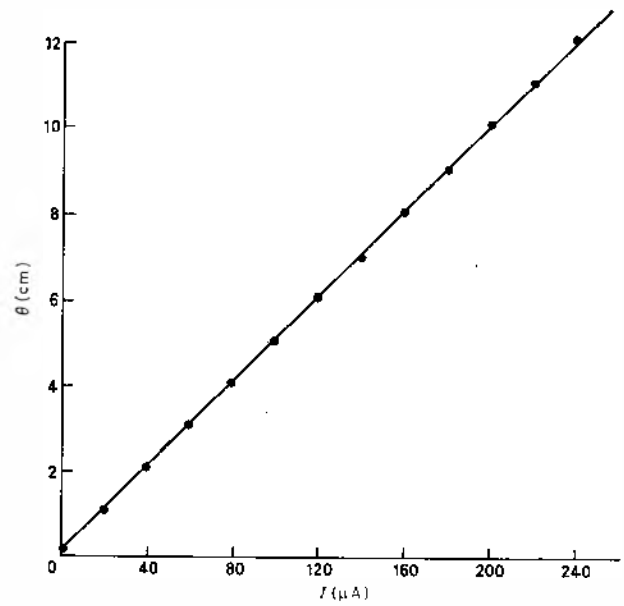
(3) Measurement of R.

(4) Measurement of shot noise as a function of diode current.

(5) Measurement of sensitivity of system.

This experiment cannot, of course, be carried out in a single three-hour laboratory period, but this is not unusual for an advanced experiment. If suitable apparatus and component values have been previously chosen then, with guidance, a student should be able to perform these measurements in three periods; the first to set up, adjust, and become familiar with the apparatus and to verify its square law response, the second to measure C, L and R, and the third to measure the shot noise and the amplifier

Fig. 4. Experimental verification of the linear relationship between $\langle v^2 \rangle$ and ***I***



sensitivity. Such a schedule does not include the time necessary for the student's task of familiarizing himself with some of the relevant theory and some background of the topic. It is therefore essential for the student to be supplied with adequate laboratory notes and suitable references.

Consideration of Errors

The uncertainty in the result quoted for e stems mainly from possible systematic calibration errors and from drifts in the amplifier gain and signal generator output. Random errors, corresponding to the scatter of points about a best straight line, have been calculated and found to be rather small by comparison. The electron charge may be expressed in terms of the measured quantities as

$$e = \frac{8\pi f_r^2 C^2 m_s}{m_f S}$$

where f_r is the circuit resonant frequency, C the capacitance, m_s the gradient of the plot of galvanometer deflexion against diode current (Fig. 4), m_f the gradient of the plot for determination of R (Fig. 3), and S the overall sensitivity of the system.

The resonance frequency could be determined to within 0·2kc/s at 363kc/s so that error from this measurement can be considered negligible. Error in the determination of C is due mainly to uncertainty in the value of the standard C_s. This was measured over a range of frequencies and by two methods, the results showing a spread of $\pm 0{\cdot}5$ per cent. This is taken to be the possible error in C. The calculated standard deviation of the gradient m_s is very small, less than 0·2 per cent, but a systematic error may have arisen due to a calibration error of the micro-ammeter. The manufacturers claim an error of less than 1 per cent, and this figure has been taken as the possible error for m_s. The calculated standard deviation of the gradient m_f is 0·6 per cent.

The greatest source of error lies in the measurement of S. The step attenuator employed was accurate to 0·1dB. This uncertainty in attenuation, together with a possible calibration error of the voltmeter, introduces an uncertainty of about 3 per cent in the d.c. calibration of the thermocouple. Drift in the amplifier gain and signal generator output (over several hours) introduces a further uncertainty in S estimated at about 1·5 per cent, giving a total uncertainty in the measurement of S about 4·5 per cent.

Combining these estimates of possible error yields a resultant of about 5 per cent for the uncertainty in the result for e. This figure is, of course, only an estimate and has not the significance of a standard deviation. The remarks in this section are mainly of interest in showing the relative contributions to the final uncertainty.

Conclusion

It has been attempted in this article to demonstrate that an experimental measurement of shot noise forms a worthwhile student experiment. A fundamental effect of considerable practical importance is illustrated, a fundamental constant is evaluated, and experience is gained in electrical measuring techniques.

As described here the experiment may be regarded as an illustration and an exercise. The topic, however, lends itself well for development as a student research project. Examples for more extended study include secondary electron emission noise and p-n junction noise. The system can be calibrated for this work by the diode shot noise. (It should be stressed here perhaps that the saturated dio has long been used by electronic engineers as a predictab noise source).

It is possible that in some laboratories electronic ge of sufficiently high quality is available for a measureme of electron charge to be made with a precision cor parable to that of the standard Millikan experiment. Eve with less elaborate apparatus the experiment remai instructive, and it forms a useful addition to the rath short list of quantitative experiments on fluctuation ar diffusion phenomena.

Appendix

Campbell's theorem[6] is concerned with the statistical stationary response of a linear system to a succession equal magnitude impulses or shocks occurring randoml This section will justify very briefly, but not at all rigo ously, the expression quoted and employed in the articl A discussion of this theorem and its conditions of validit applications, and further references are given by Ma Donald[4].

Suppose impulses of very short duration arrive random in time at a system at an average rate N per second, an suppose the response of the system to a single impulse $f(t)$. Then the resultant response at time t of the syste is $V(t)$ given by

$$V(t) = \sum_{-\infty}^{t} f(t - t_r)\, p_r$$

where $p_r = 0$ or 1 and represents the probability of a impulse occurring during an interval δt_r at time t_r. It h been assumed that a sufficiently fine time division can b made such that the probability of two or more impuls arriving during an interval δt_r is negligible. Denotin average values (over an assembly of systems) by $\langle\rangle$, the $\langle p_r \rangle = N\,\delta t_r$.

Now the mean square deviation $\langle v^2 \rangle$ of $V(t)$ is define by the relationship $\langle v^2 \rangle = \langle V^2(t) \rangle - \langle V(t) \rangle^2$. One ma therefore proceed to calculate expressions for the tw terms in this expression.

Thus $\langle V^2(t) \rangle = \sum_{t_r} f^2(t - t_r)\langle p_r \rangle + \sum_{t_r \neq t_s} f(t - t_r)\, f(t - t_s)\langle p_r p$

and $\langle V(t) \rangle^2 = \sum_{t_r} f^2(t - t_r)\,\langle p_r \rangle^2 + \sum_{t_r \neq t_s} f(t - t_r)\, f(t - t_s)\,\langle p_r \rangle \langle p$

Now since the individual events are completely u correlated $\langle p_r p_s \rangle = \langle p_r \rangle \langle p_s \rangle$

Further $\langle p_r \rangle^2$ becomes negligibly small compared wit $\langle p_r \rangle$ as δt proceeds to the limit 0.

Hence $\langle v^2 \rangle \simeq \sum f^2(t - t_r)\langle p_r \rangle = N \sum f^2(t - t_r)\,\delta t_r$

In the limit $\langle v^2 \rangle = N\int_{-\infty}^{t} f^2(t - t_r)dt_r = -N\int_{\infty}^{0} f^2(x)dx$

or

$$\langle v^2 \rangle = N\int_{0}^{\infty} f^2(t)dt$$

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The Non-linearity of Fixed Resistors

By P. L. Kirby*

The use of a voltage coefficient in defining the magnitude of the non-linearity of fixed resistors has serious deficiencies and an alternative method is proposed, involving the measurement of third harmonic distortion. In a simple case, a third harmonic output is produced, proportional to the cube of an applied fundamental voltage and on this basis a new unit entitled 'Third Harmonic Index' (t.h.i.) is derived. An explanation is suggested for the origin of non-linearity and a reason given for its correlation with current noise. Typical values of t.h.i. are quoted for film resistors over a wide range of values and future applications include its use in the evaluation of basic resistor performance, for test specification purposes and as a method of screening to ensure very high reliability.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 781)

MANY electronic components in the form of two-terminal devices are required to exhibit a perfectly ıear relationship between two dependent variables. ҳamples of this include resistors (current and voltage), ;pacitors (charge and voltage) and inductors (rate of ıange of current and voltage). Any divergence from such strictly linear relationship is usually regarded as a fault ›ndition and is therefore to be avoided.

There are generally two ways in which 'non-linearity' akes itself apparent in supposedly linear components. ı the first place at some high level of one of the vari-›les there can be a sudden onset of non-linearity such ; would occur in a 'breakdown' condition. Secondly, ıere may be a small but consistent deviation from strict nearity present at all levels of use as, for example, occurs hen a resistor exhibits some slight departure from Ohm's ıw. It is this latter type of imperfection which will be iscussed.

In the case of fixed resistors non-linearity can be escribed in terms of an apparent voltage coefficient. It is ot surprising that, as a defect, this term has often been uoted in a context together with other non-perfect para-ıeters such as temperature coefficient, instability or drift, urrent noise etc. Indeed, at a time when component manu-ıcturers are striving hard to increase the reliability of ıeir products is it appropriate that the maximum possible ttention should be directed towards all aspects in which ıe component falls short of perfect or ideal behaviour.

Not all types of fixed resistor exhibit non-linearity to a ıeasurable degree. Resistors which utilize a metallic con-uctor, say in the form of a wire, exhibit neither ohmic on-linearity nor current noise over the range of measure-ıent permitted by available apparatus. In contrast, esistors utilizing a semiconducting material either in a ɔlid form of construction or as a film on an insulating ɔrmer do manifest both non-linearity and current noise. 'his is an indication that both of these phenomena, whose pparent relationship is discussed later, are related to the ıature of the resistive material concerned.

The non-linearity of the current/voltage characteristic n any of the usual types of fixed resistor is extremely small nd cannot be detected in a simple linear graphical pre-entation of current versus voltage. The effect can be ›bserved using sensitive apparatus and is sometimes ıssessed by a direct measurement of the 'voltage co-fficient' of resistance. This approach suffers from certain lisadvantages. The normal procedure involves the rapid neasurement of resistance at two levels of voltage. The nethod cannot be applied to film resistors due to the nevitable heating effect of the element and the variation n resistance which results from the effect of the temperature coefficient. An improved system involving the application of power for only 0·2sec was utilized in a commercial apparatus for the measurement of voltage coefficient. This, too, was found to be more sensitive to the temperature coefficient of film resistors than to their non-linearity.

A further serious disadvantage of a 'voltage coefficient'

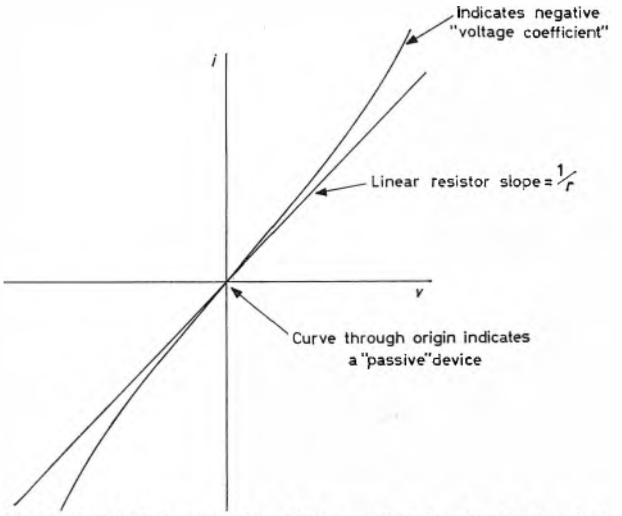


Fig. 1 Graphical representation of the non-linearity of a fixed resistor

relates to the very considerable dependence of this parameter on the level of applied voltage at which it is measured. If the relationship between current and voltage departs from perfect linearity, then it is hardly to be expected that the 'resistance' (derived from the slope of the i/v curve) will, in turn, vary linearly with the voltage and provide a 'voltage coefficient' which is constant at all levels of applied voltage.

An alternative indication of the non-linearity of a resistor may be obtained from the measurement of the amplitude of harmonics generated in the resistor when a pure fundamental sine wave signal is applied. Such harmonics can be detected by the use of sensitive apparatus[1] and commercial equipment has been developed to measure the ratio of the magnitude of the third harmonic component to that of the applied fundamental signal.

The small amount of non-linearity which is usually found in fixed resistors can be represented by a mathematical expression of the following type:

$$i = \alpha v + \beta v^3 + \gamma v^5 + \delta v^7 + \ldots \quad \ldots\ldots\ldots (1)$$

This expression utilizes only the odd powers of a polynomial series and represents a device which is 'symmetrical' and 'passive' in the electrical sense (Fig. 1). The deviation from perfect linearity is usually very small

* Welwyn Electric Ltd.

and coefficients of increasing order in the above expression rapidly decrease in magnitude. It is found that, in all normal cases, the voltage coefficient is negative and this implies that the sign of β representing the third order coefficient is negative. The effect of a device having this characteristic on a pure a.c. waveform will now be considered. Due to the symmetry of the above expression, the output will contain a number of *odd* harmonics. An indication of the magnitude and phase relationship of the harmonics is given by the Fourier series derived in Table 1. It is, of course, most important to distinguish between a *coefficient* which represents the magnitude of the terms of increasing order in the polynomial expression representing current/voltage characterestic and the *amplitude* of the harmonics of ascending order. Nevertheless, an interesting practical relatonship between the two does emerge. In the first place the expression for the amplitude of the n^{th} harmonic includes factors relating only to the n^{th} *and higher order* coefficients of the original expression. Furthermore, as the n^{th} coefficient is likely to be considerably greater than the next higher coefficient it is likely that the magnitude of the n^{th} harmonic will largely be governed by the magnitude of the n^{th} order coefficient. For example, Millard[1] showed that, with an $\frac{1}{4}$W carbon film resistor at all levels up to 0·75W, the third harmonic could be attributed solely to the third order term of the original expression. The equations suggest that such a simplification will not be valid if the applied voltage is large, as this will augment the effect of the higher order coefficients.

If, for simplification, it is assumed that fifth and higher order curvature is negligible, then the current/voltage relationship is represented by a cubic expression. It is then found that the amplitude of the third harmonic voltage is proportional to the third power of the amplitude of

TABLE 1

Non-Linearity of Current/Voltage Characteristic

$$i = \alpha v + \beta v^3 + \gamma v^5 + \delta v^7 + \ldots$$

When $v = V\sin\omega t$

then $I = \alpha V\sin\omega t + \beta V^3\sin^3\omega t + \gamma V^5\sin^5\omega t + \delta V^7\sin^7\omega t + \ldots$

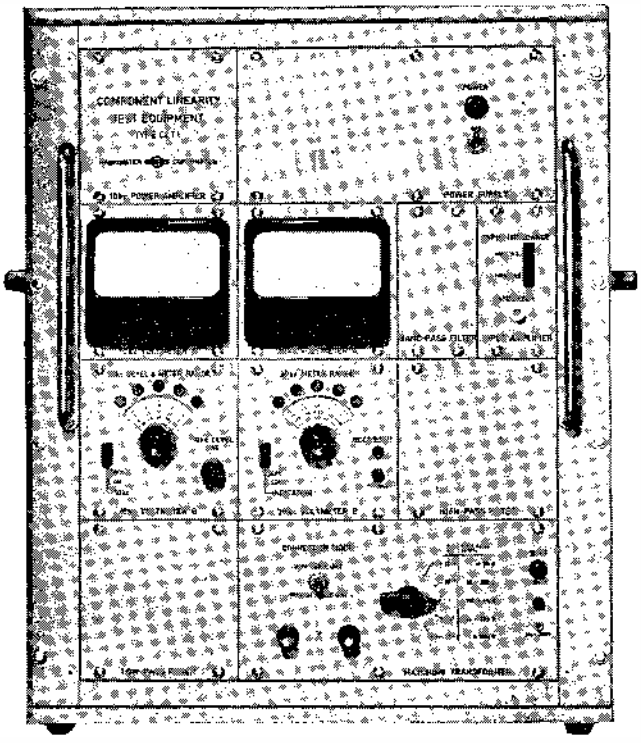
or in terms of fundamental 3rd harmonic 5th harmonic 7th harmoni

$$\begin{aligned}
I = \;& \alpha V\sin\omega t \\
& + \beta V^3\left(\tfrac{3}{4}\sin\omega t - \tfrac{1}{4}\sin 3\,\omega t\right) \\
& + \gamma V^5\left(\tfrac{5}{8}\sin\omega t - \tfrac{5}{16}\sin 3\,\omega t + \tfrac{1}{16}\sin 5\,\omega t\right) \\
& + \delta V^7\left(\tfrac{35}{64}\sin\omega t - \tfrac{21}{64}\sin 3\,\omega t + \tfrac{7}{64}\sin 5\,\omega t - \tfrac{1}{64}\sin 7\,\omega t\right)
\end{aligned}$$

Fourier Series for Symmetrical Non-Linearity up to the Seventh Order

$$\begin{aligned}
I = & \left(\alpha V + \tfrac{3}{4}\beta V^3 + \tfrac{5}{8}\gamma V^5 + \tfrac{35}{64}\delta V^7\right)\sin\omega t. \\
& - \left(\tfrac{1}{4}\beta V^3 + \tfrac{5}{16}\gamma V^5 + \tfrac{21}{64}\delta V^7\right)\sin 3\omega t. \\
& + \left(\tfrac{1}{16}\gamma V^5 + \tfrac{7}{64}\delta V^7\right)\sin 5\omega t. \\
& - \left(\tfrac{1}{64}\delta V^7\right)\sin 7\omega t.
\end{aligned}$$

Fig. 2. Commercially available test set for the measurement of third harmonic distortion



the fundamental applied waveform. At low powers it been found that this relationship holds for fixed f resistors and solid composition resistors[2]. At their norr ratings the latter produce third harmonic voltages dep dent more nearly on the square of the applied fun mental voltage. Film resistors of a wide range of valu tested at nominal rating show that the lower values m rigorously obey the cubic relationship, whereas hig values tend in some cases towards a power law betw the square and the cubic forms. This suggests that in latter cases, where high test voltages are applied, efficients of a higher order than the third are playing significant part in the current/voltage relationship.

A cubic expression is, however, a very useful appro mation for all cases. A suitable unit for measuring the tortion of the device is given by expressing the ratio the third harmonic, in microvolts, to the cube of applied fundamental in volts (both in r.m.s.). In line w accepted practice for the expression of the current no of a resistor, in terms[3] of a 'Noise Index' (n.i.) whe

$$n.i. = 20\log\frac{r.m.s.\ microvolts\ noise\ in\ a\ frequency\ decade}{d.c.\ volts\ applied}$$

a similar unit has been derived to permit more conveni analysis of the measurement of distortion.

Third Harmonic Index (t.h.i.) =

$$20\log\frac{r.m.s.\ microvolts\ of\ third\ harmonic}{(r.m.s.\ volts\ of\ applied\ fundamental)^3}$$

It is found that, in the range of fixed resistors of a vari of different types and values, this unit ranges from ab −80 to −20dB.

The level of non-linearity varies considerably in a ba of nominally identical resistors. The use of a logarithn unit is advantageous as the magnitude of the t.h.i. appro mates to a normal distribution. This permits the use statistical parameters and facilitates the application control to this particular variable. Table 2 illustrates series of grouped frequency tables for the t.h.i. of resistors of different types and values. Largely due to use of the 'cube of the applied voltage' in the definiti of t.h.i. it is found that this parameter *decreases* w increasing resistance value. This contrasts with the no index which usually *increases* with resistance value. In a batch of resistors the magnitude of both t.h.i. and n.i. ha remarkably similar distributions. Both are approximat normal when such logarithmic units are used, but f

TABLE 2
DISTRIBUTION OF T.H.I. FOR ½ WATT FILM RESISTORS OF VARIOUS VALUES

VALUE	10Ω		100Ω		1kΩ		10kΩ		100kΩ		1MΩ	
RESISTOR TYPE	OXIDE	CARBON	OXIDE	CARBON	OXIDE	CARBON	OXIDE	CARBON	OXIDE	CARBON	OXIDE	CARBON
T.H.I. (dB)												
+10 to + 5·1	2		1									
+ 5 + 0·1	98	2										
0 − 4·9		85										
− 5 − 9·9		13										
−10 − 14·9			2									
−15 −19·9			1									
−20 −24·9			12	4								
−25 −29·9			47	8								
−30 −34·9			22	48		4						
−35 −39·9			12	40	1	5	1	1				
−40 −44·9			3		6	11	2	18				
−45 −49·9					14	15	9	81	1			
−50 −54·9					19	41	11		2	3		
−55 −59·9					34	24	17		2	13		
−60 −64·9					21		56		4	23		6
−65 −69·9					5		4		10	36		63
−70 −74·9									24	22		31
−75 −79·9									48	2		
−80 −84·9									9			

ently indicate some skewness which incorporates a 'tail' ending into the higher levels.

The value of the t.h.i. of any resistor, unlike its 'voltage efficient', is practically independent of the level of plied test voltage. Most investigations of the harmonic ntent of the distortion introduced by the non-linearity a fixed resistor suggest that the second harmonic and l harmonics higher than the third are small compared h the latter. Thus for most purposes including the itine checking of the performance of resistors, a asurement of the third harmonic ratio and its expres-n as a third harmonic index is quite adequate. Commer-l apparatus has been developed for the measurement of tortion in terms of the magnitude of the third harmonic nponent (Fig. 2).

f a more detailed investigation of the behaviour of the istor is required it may be found that this method of essing non-linearity is not entirely free from inter-ence due to the temperature sensitivity of the resistor nent. The self heating of the element in association h the temperature coefficient can produce a sym-trical non-linearity between current and voltage which ılts in a polynomial expression utilizing only the odd vers as in equation (1). In this case, however, there uld be some slight difference in phase between the monic voltages due to true ohmic non-linearity and se resulting from the heating effect. For a thorough estigation apparatus would be required which could measure the amplitude of higher order harmonics and give some indication of their relative phase.

It was noted above that only those types of fixed resistor which incorporate a semiconducting material as the negative element exhibit ohmic non-linearity or current noise. A simplified explanation can be given concerning the basic physical mechanisms which give rise to these two effects in the materials concerned. There is as yet no generally accepted complete explanation for the occurrence of excess or current noise but most theories have centred around the modulation in conductivity which arises when current carriers pass into and out of the conduction bands. A similar general picture cannot be utilized to explain non-linearity as there is no reason why any such interchange of carriers should be dependent upon the magnitude of the applied voltage. Those regions where the energy levels are disturbed, as may occur due to the existence of surface states or at 'junctions' within the material, may be regarded both as a potential source of current noise and of non-linearity or variation of conductivity with applied potential. In other words, while current noise could occur in a semiconducting material without there necessarily being any form of discontinuity or junction, non-linearity can only occur in regions where some form of potential barrier has been produced.

Although a junction between two regions of the material, where different energy levels or different levels of carrier concentration exist, may clearly involve ohmic non-

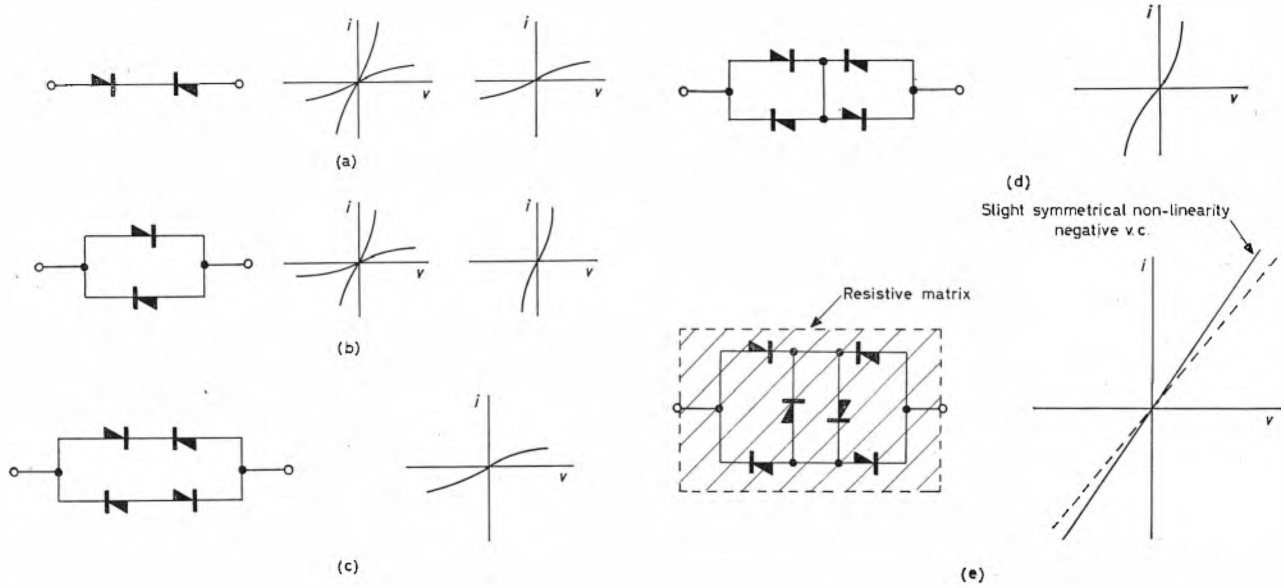


Fig. 3. Combination of rectifying junctions to synthesize the response of a fixed resistor

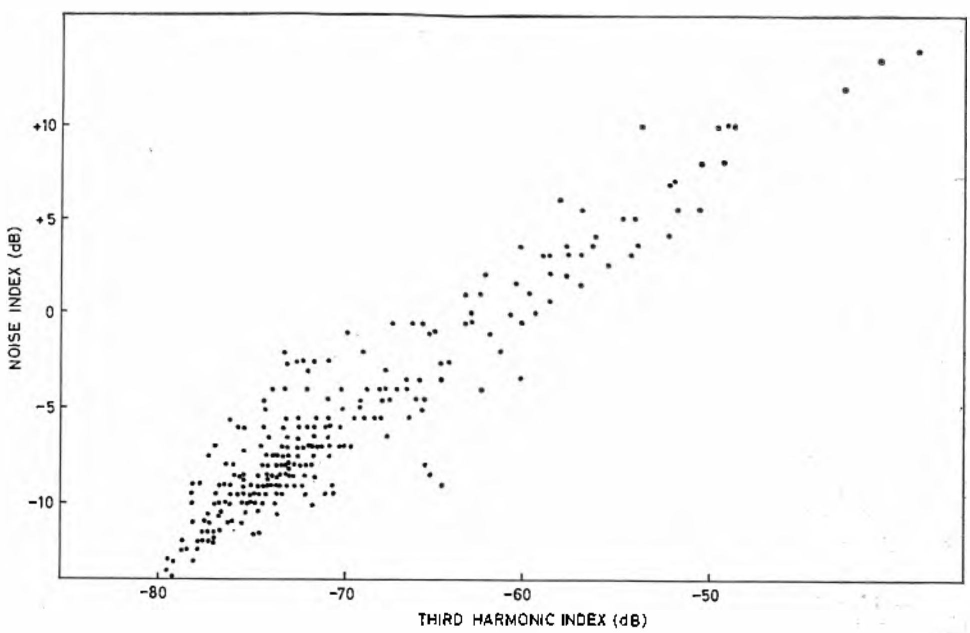
linearity, such a single junction will have an assymmetrical characteristic. This will occur whether the 'junction' occurs between p and n material or n and n+ material or between any regions where the energy levels or carrier concentrations differ. Fixed resistors, in contrast, exhibit a small degree of non-linearity of a symmetrical form. The resistive element consists of an amorphous or fine-grain polycrystalline aggregate of the semiconducting material concerned. In such a material the 'junctions' which may occur are likely to be widely distributed both in regard to the magnitude of the potential barriers and also with regard to the position and direction which they occupy in the electrical matrix. The resulting series-parallel combination of a wide variety of junctions can be shown to result in a slightly non-linear characteristic but one which, due to the even chance of a barrier appearing in one direction or in the other, has a reasonably symmetrical form. This simplified picture suggests that the basic source of ohmic non-linearity is the complex random set of quasi-junctions which are distributed throughout the resistive material.

The illustrations given in Fig. 3 show how a system of simple rectifying junctions can be built up to give an assembly whose response begins to approximate to the *i/v* characteristics of a resistive film. Figs. 3(a) and 3(b) represent two opposite junctions connected respectively in series and parallel. The former gives a high impedance unit with a positive voltage coefficient and the latter a low impedance unit with a negative coefficient. Similar differences in characteristics are found between Figs. 3(c) and 3(d) and here it can be seen that the provision of a cross link produces an all-important difference in the response of the assembly. Bearing in mind that the resistive film being simulated has a very low order of non-linearity but that it does invariably have a negative voltage coefficient, one must conclude that the 'cross-links' in the resistive film, which must certainly exist in practice, are of very great importance.

Fig. 3(e) illustrates a more complex assembly of rectifying junctions having the required negative coefficient and if this is now imagined to incorporate many additional resistor cross links, or preferably to be virtually immersed in a continuous resistive matrix, then the *i/v* characteristic is further modified and approaches the known behaviour of actual fixed resistors. A qualitative explanation of non-linearity of this nature can be utilized to explain the existence of ohmic non-linearity in both solid composition and film resistors.

The magnitude of the current noise of a film resistor

Fig. 4 The relationship between third harmonic distortion and current noise in a group of over 400 cracked-carbon resistors



and its dependence on the thickness, width and length of the resistor track has been studied previously[4]. It appears that the observed facts can be explained if it is assumed that current noise is produced at a large number of discrete sources distributed throughout the material of the film. The behaviour of the system can then be analysed in terms of a number of incoherent noise generators in a conducting matrix. The series and parallel combinations of random generators coupled with the series resistance and shunting effects of the matrix suffice to explain the observed effects.

The generation of noise throughout all parts of the material of the film due to the transition of carriers to and from the conduction band must still be regarded as a possible explanation of the phenomenon. Nevertheless, this picture of discrete random noise generators is very similar to the idea of a distribution of the 'junctions' which can be the source of non-linearity. The statistical relationship between current noise and non-linearity lends some emphasis to the idea that the major source of noise may correspond to the sole source of non-linearity at quasi-junctions dispersed throughout the film.

It would therefore be expected that there will be a correlation between the current noise and the ohmic non-linearity of a fixed resistor. A great deal of evidence has been compiled to illustrate this correlation and Fig. 4 shows the relationship between these two parameters in a group of over 400 ¼W 500kΩ carbon film resistors where the correlation coefficient was found to be 0·90.

In the case of a film resistor both phenomena can be affected by similar geometrical factors. Similarly, if an irregularity occurs at some point on the film then both the current noise and the non-linearity may be increased above the level associated with similar resistors without such a defect. The latter may be a possible subsequent cause of inferior performance and a preliminary measurement of noise or non-linearity may permit the elimination of this potential failure from the batch of resistors. Such a procedure must not be regarded as a complete guarantee against resistor failure as may sometimes have been suggested, but does go some way towards improving the reliability of the product. The choice as to whether current noise or non-linearity should be used for this purpose depends on particular circumstances. An important criterion is the reliability with which the instrument will operate on an automatic production line in carrying out measurements rapidly without interference from neighbouring equipment. It now appears that non-linearity test sets (Fig. 2) may well prove to be preferable to noise test sets in these respects.

Conclusion

The measurement of distortion given by the magnitude of the third harmonic component has been found to be a most convenient method of assessing the ohmic non-linearity of fixed resistors. Over a wide range of resistance values and at different levels of applied voltage it gives a repeatable measurement which is less affected by other resistor parameters than a measurement of so-called voltage coefficient. The latter parameter may well have lost some of its original popularity for reasons which have been explained. It is likely that a measurement of non-linearity by reference to the third harmonic distortion particularly when expressed in the form of a 'third harmonic index' will find greater future application. Such techniques have considerable value in laboratory investigations into resistor quality, and versatile apparatus of high sensitivity with the possibility of measuring the 3rd and 5th harmonic and their phase relative to the applied fundamental would be useful. In addition, however, there is the need for reliable lower-priced production line equipment to measure rapidly and reject automatically any resistor having a level of non-linearity greater than a prescribed limit. Finally, there is the likely future use of third harmonic distortion, perhaps in the form of the new parameter t.h.i. in resistor specifications. This could replace previous attempts to measure a voltage coefficient and would be complimentary to the growing use of a noise index with which the t.h.i. has such a significant correlation. An indication of the future areas of application of distortion measuring apparatus is given in Table 3.

TABLE 3

Third-Harmonic Distortion Measurements—Future Applications

(i)	Laboratory investigations into resistor quality. Measurement of *i/v* symmetry, 3rd & 5th harmonics, and relative phase to distinguish t.c. etc. Versatile apparatus of high sensitivity.
(ii)	Component specifications,—replacing "voltage coefficient." Use of t.h.i. complimentary to noise index. Reliable and repeatable measurements under known conditions.
(iii)	Automatic production test,—rejection at given level. Data accumulation (on log. scale, i.e. t.h.i.) for process control. Robust lower-price apparatus—rapid measurement free from interference.

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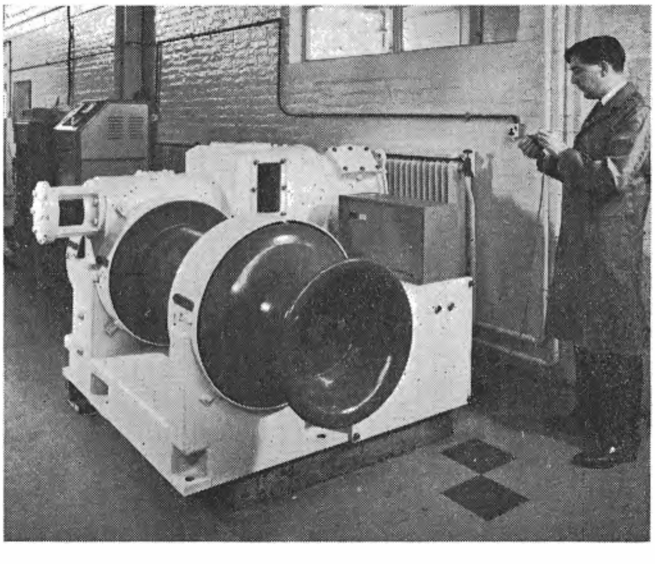
S.C.R. Motor Control

The recently established electronics division of Clarke, Chapman and Co. Ltd, of Newcastle upon Tyne, has produced a series of silicon controlled rectifier circuits for speed control of motors in marine applications.

The photograph shows a marine winch driven by a 50 h.p. motor, the speed of which is controlled from creep to maximum with constant torque by an s.c.r. circuit.



A 1kW Invertor for Mains Frequency Operation

By W. D. Gilmour*

An invertor, delivering up to 1kW at accurately known frequencies between 45 and 65c/s, and with good stability of output voltage, is described. The output is sinusoidal at loadings of up to 300W becoming progressively more square as the loadings are increased. A low voltage control system is used

(Voir page 776 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

FOR testing magnetic recording tape decks, an a.c. supply of greater frequency stability than the normal supply mains is sometimes required for measurements of wow, flutter, and the slip occurring between the capstan and the recording tape. The waveform of the supply is not of primary importance. Also, for testing equipment for export, an a.c. supply, of good waveform, stable in voltage and frequency, and capable of adjustment between 45 and 65c/s, is required. The invertor to be described serves as a supply for both purposes, giving a sinusoidal output at loadings of up to 300W, and more square waveforms at higher loadings up to its maximum rating of 1kW.

The invertor is of basically conventional design, wherein input power from the mains is rectified and smoothed, and is then switched alternately in the two balanced halves of an output transformer, across the secondary of which is developed the output waveform. For the switches, although mechanical devices could be used, electronic operation is far more convenient, and the choice lay between thyristors and mercury vapour or inert gas thyratrons. Because the loads can reflect considerable inductance into the primary of the output transformer, the back e.m.f.'s to be withstood by the switches are rather high for existing thyristors of moderate cost. The long warming-up time of mercury thyratrons was considered a dis-

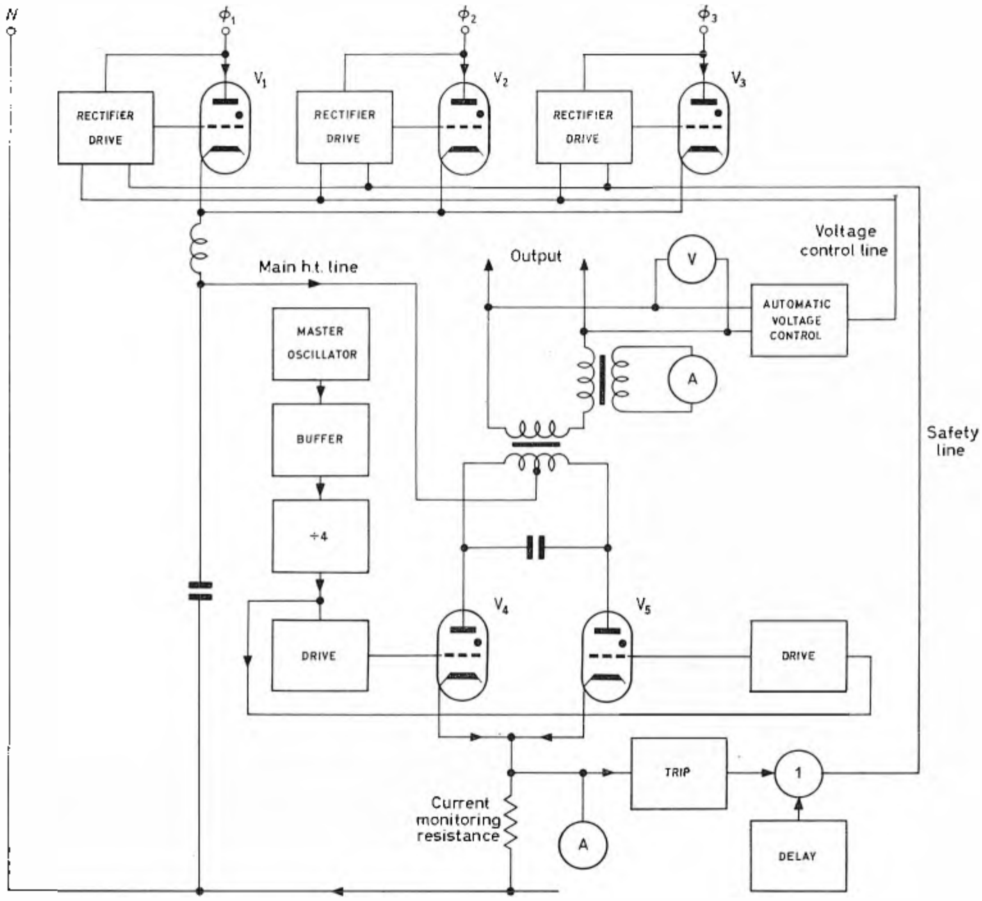


Fig. 1. Arrangement of the equipment

* *E.M.I. Electronics Ltd.*

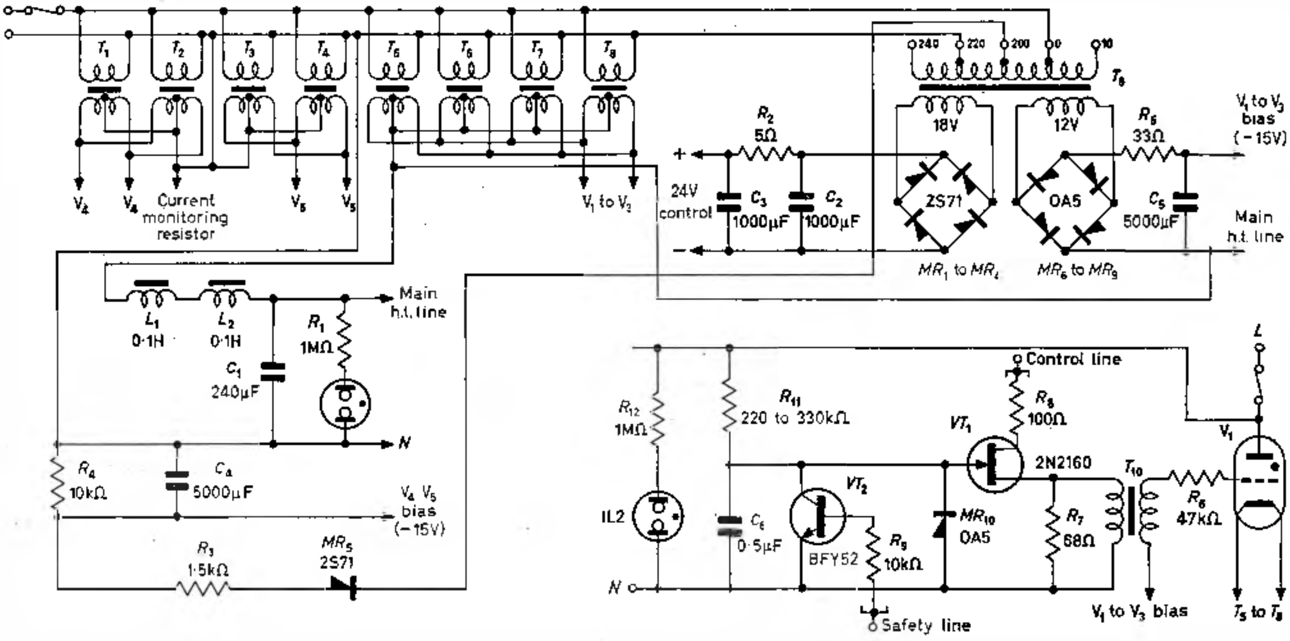


Fig. 2. The main power supplies

ıvantage, and therefore inert gas thyratrons were chosen ; the switching means, and the type finally decided upon as the XR1-3200. This valve was also chosen for the put rectifiers, but proved somewhat liable to arcing ıck in this position, and all valves were finally changed ı the type XR1-3200A, and have proved extremely ıccessful.

Although solid state devices were not used in the main ırrent carrying circuits, the great convenience of semi- ›nductors has been used throughout the control circuits, hich all operate from a +24V supply.

;eneral Principles

Fig. 1 shows the arrangement of the whole equipment. three phase 50c/s mains supply is rectified by the half- ave thyratron circuit, V_{1-3}. The voltage of the main .t. supply to the invertor proper, and hence the ouput oltage of the latter, is controlled by adjusting the firing ngle of these valves. Under fault conditions the rectifier ıyratrons can be blanked off, thus interrupting the supply › the invertor without any mechanical contactor. A ormal choke input filter is used to smooth the h.t. supply. he frequency of the invertor is set by a master oscillator ınning at four times the desired output frequency to void any pulling effects from 50c/s hum. The master scillator is stable to ±0·1 per cent in frequency, but rovision is made for the use of a more precise external scillator if required. After buffering, the output from the scillator is divided by four in two conventional binary ividing stages, and is then used to drive the invertor ıyratrons. Should the invertor be overloaded, or fail › operate correctly, it will draw an abnormally large ırrent, which is detected by the voltage developed across current monitoring resistance, and this operates an utomatic trip circuit that biases off the rectifier thyra- ons, as described above. A multivibrator circuit, which ın be optionally switched into a monostable or istable configuration, re-applies the h.t. supply after a short delay, or causes the main h.t. to remain switched off, depending on the switch setting selected. For some applications an immediate resumption of the output is desirable, in others a positive indication that a 'hiccup' has occurred is required. The output voltage is automatically controlled to within ±2V by a circuit which adjusts the firing angle of the rectifier thyratrons. The necessary warming-up time for the thyratrons is measured electronically, and until this time has elapsed, no h.t. can be drawn from the circuit.

Detailed Description

Main Power Supplies

Fig. 2 shows the main power supplies. Diodes MR_{1-4} form a full wave bridge rectifier to supply +24V for the control circuits. Diodes, MR_{6-9} form a similar circuit for biasing the rectifier thyratrons, the whole circuit resting at rectifier cathode (h.t.) potential. Standing bias for the invertor thyratrons is generated in the half-wave rectifier MR_5. The very large smoothing capacitors are necessary to ensure that no thyratron can strike after a short interruption in the mains supply, when the cathodes are still hot. The control circuit for one phase of the main rectifier is also shown. The input voltage is delayed by 90° in the network R_{11} and C_6. The negative half cycle is removed by the diode, MR_{10}, but the positive half cycle is applied to the emitter of the unijunction transistor VT_1, provided that the safety line is earthy. If the latter is positive, VT_2 will conduct and no potential will be applied to VT_1. When the emitter potential exceeds the firing potential of VT_1, C_6 will be discharged into the low impedance primary winding of T_{10}, and the large positive pulse produced from this step-up transformer will fire the thyratron, V_1 which will be extinguished when its anode potential becomes negative with respect to its cathode. The point at which firing occurs on the positive half cycle of anode voltage will be determined by the voltage on the control line, which, like the safety line, is common to all three phases. The more positive this voltage, the

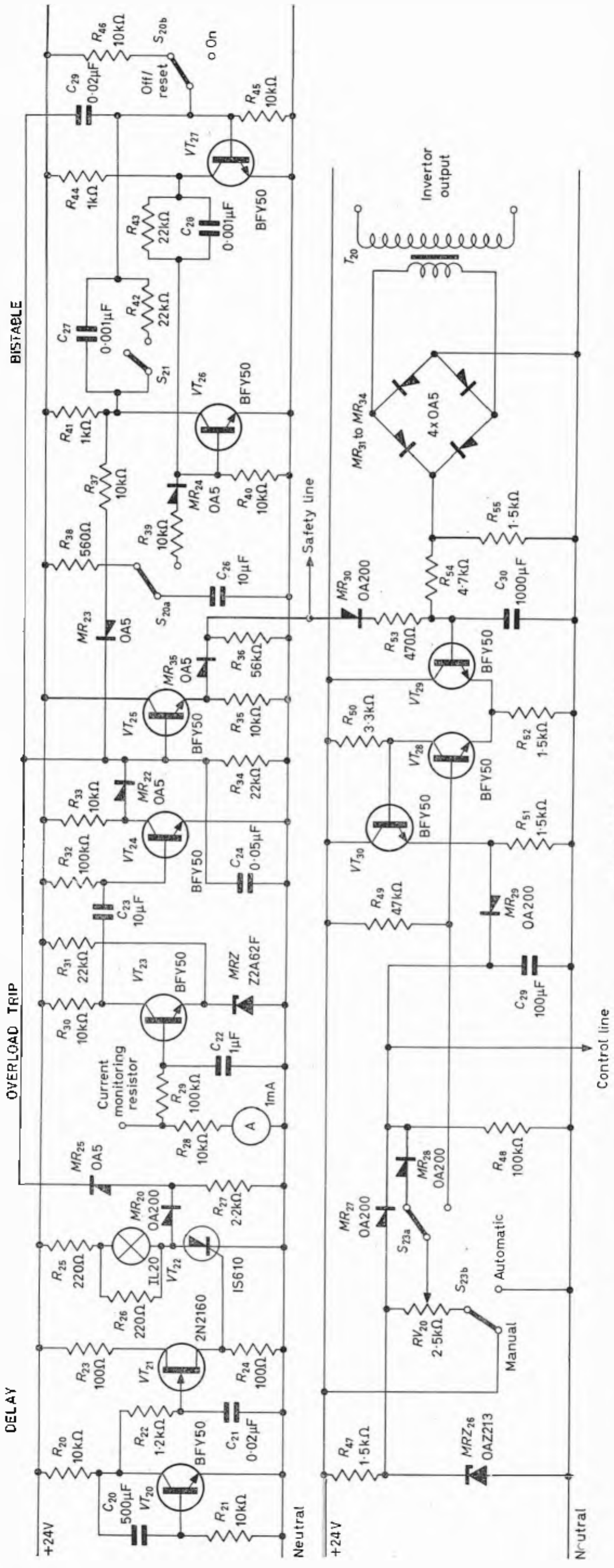


Fig. 3. The control circuits

later will firing occur, for a unijunction transistor will break down when the emitter is more positive than a fixed fraction of the interbase voltage. The unijunctions used were selected to be reasonably well matched in characteristics, and a final adjustment to obtain an even distribution of current over the three rectifier thyratrons may be made by adjusting R_{11}. A fuse protects the thyratron in the event of arcing back, but all other faults are taken care of by the safety line rapidly going positive, thus removing the drive from all the rectifiers. The integrity of the fuse and the presence of h.t. are indicated by neon lamps.

Control Circuits (Fig. 3)

(a) *Warming-up Timer:* The initial 60sec (minimum) warming-up time, required by the thyratron filaments, is measured by the Miller integrator circuit, VT_{20}. Initially the collector of this stage is earthy, but it rises in voltage linearly as C_{20} charges in the usual way. When the breakdown voltage of the unijunction transistor, VT_{21}, is exceeded, a firing pulse is passed to the thyristor, VT_{22}, which breaks down, thus lighting the 'ready' lamp, IL20, and removing the positive bias supplied through MR_{20}, MR_{25}, VT_{25}, and MR_{35} to the safety line.

(b) *Overload Trip and ON/OFF Switching*: After VT_{22} has fired, if S_{20} is placed to 'on', the positive pulse derived from discharging C_{26} into the base circuit of VT_{26}, will cause this transistor to conduct, thus removing the positive bias previously applied through MR_{23}, VT_{25}, and MR_{35} to the safety line. The rectifier thyratrons can now fire and the invertor will operate. VT_{26} forms one half of a multivibrator with VT_{27}. If S_{21} is closed the pair operate as a bistable circuit, if S_{21} is open as a monostable. VT_{26} will thus remain conducting so long as VT_{27} is cut off. VT_{27} may be made to conduct by operating the on/off switch, S_{20}, or by a positive pulse applied through C_{29} from the overload circuit, which will now be described. The input from the current monitoring resistor is smoothed by R_{29} and C_{22} and applied to the base of VT_{23}, which, however, will not conduct until its base becomes positive to its emitter potential set by the Zener diode, MRZ_{21}. Under normal operation this never occurs, but should an excessive current be drawn, a negative pulse will be developed at the collector of VT_{23}, thus cutting off the normally conducing VT_{24}, and applying a positive pulse to the safety line, thus cutting off the rectifier thyratrons and removing h.t. The same positive pulse is also applied to the base of VT_{27}, thus altering the state of the multivibrator VT_{26}, VT_{27} as described above. Depending on the setting of the switch S_{21}, h.t. will or will not be restored automatically without re-operating S_{20}.

(c) *Voltage Control:* The Zener diode MRZ_{26} provides a stable reference voltage of +12V with respect to the negative control rail. Under manual control this merely limits the negative excursion of the control line which

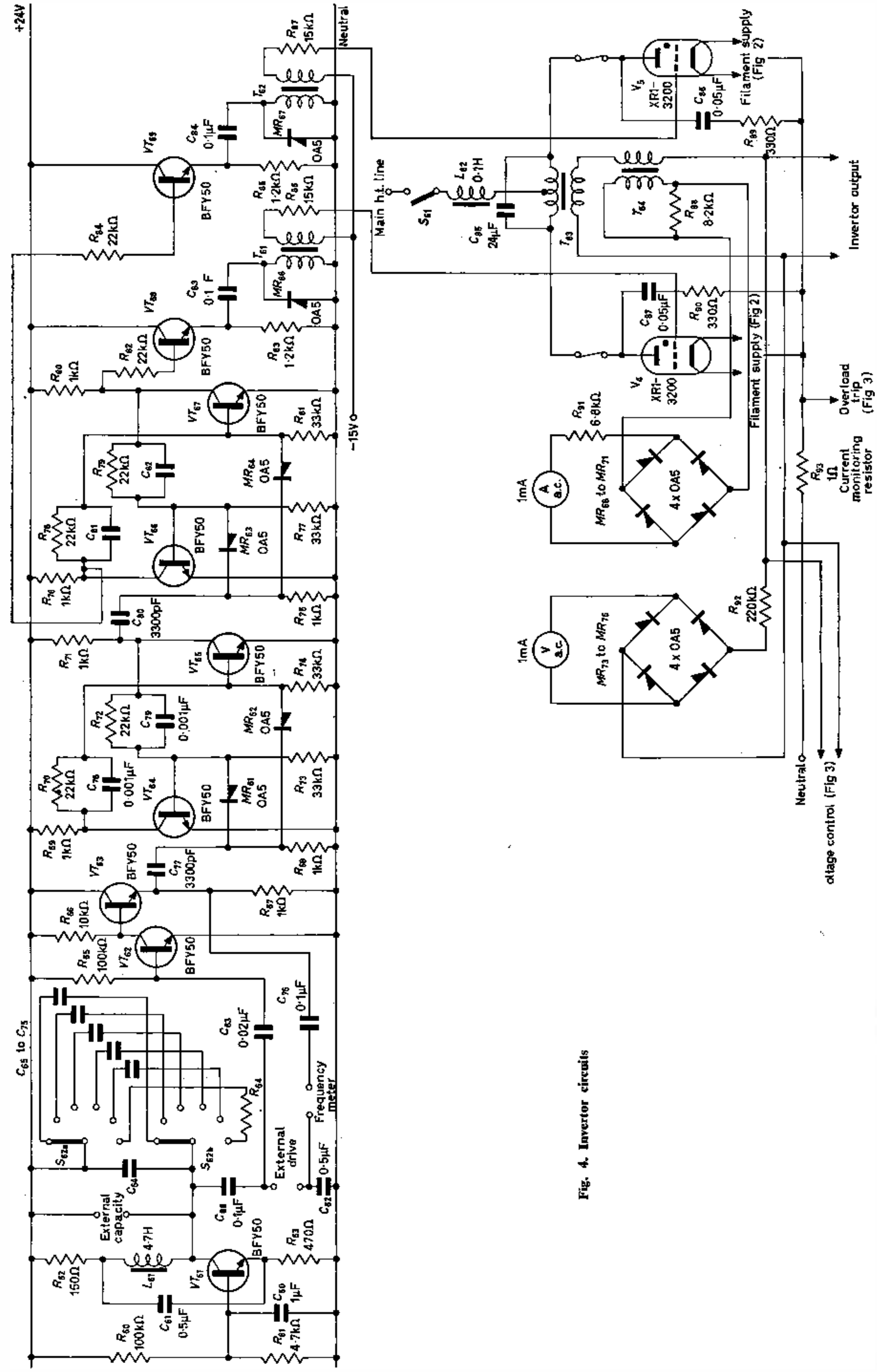


Fig. 4. Inverter circuits

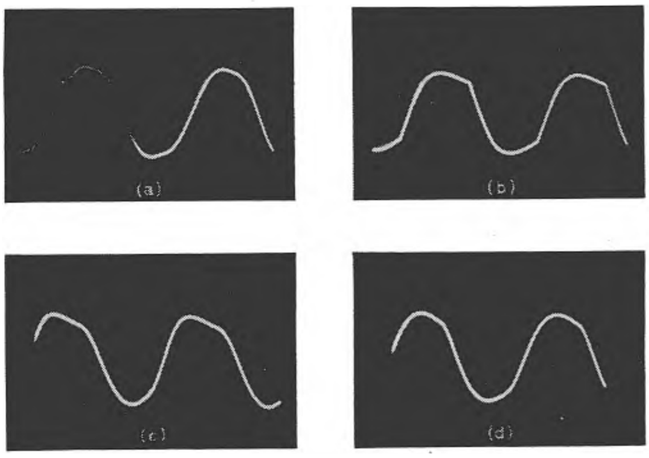


Fig. 5. Output waveforms with various loads
(a) 0A, 250V, 50c/s
(b) 2A resistive, 250V, 50c/s
(c) 0·5A television receiver (half-wave rectifier), 250V, 50c/s
(d) 1·2A tape deck, 250V, 50c/s

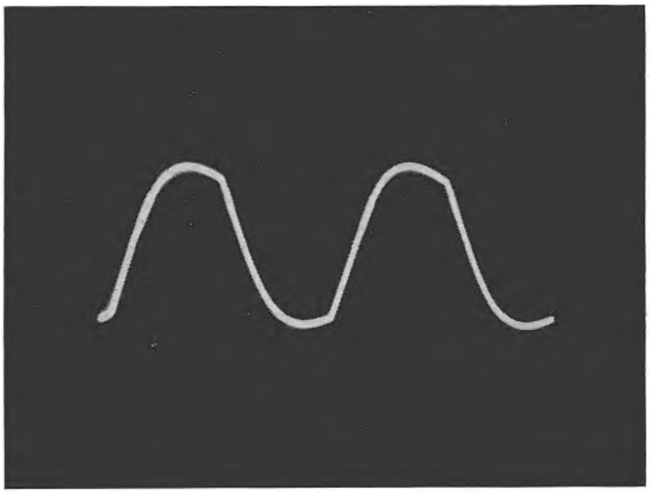


Fig. 6. Output with 2A resistive load, 250V, 60c/s

can be adjusted by RV_{20} from +24V (rectifiers almost fully off) to +12V (fully on). Under automatic control, RV_{20} is switched to provide a reference voltage to the base of transistor VT_{28}, forming one half, with VT_{29}, of a long-tailed pair, wherein the reference voltage is compared with the average value of the output voltage of the invertor, rectified in the full wave bridge MR_{31-34}. The output from the collector of VT_{28} is passed through an emitter-follower, VT_{30}, to the control line. Under manual control R_{49} causes VT_{28} to conduct heavily and thus this output cannot contribute, because of the diode OR gating, which will select the most positive source, to the voltage on the control line. Under automatic control, C_{30} is initially charged positively from the safety line, before the invertor is switched on. When it is switched on the safety line goes to earth potential and C_{30} discharges until its voltage falls to the rectified output from MR_{31-34}, when the normal stabilizing action occurs. This circuit ensures that the voltage output of the invertor is initially low (about 50 to 100V, depending on the load applied) and gradually builds up to the controlled value (normally 240V).

INVERTOR CIRCUITS (Fig. 4)

The master oscillator stage, VT_{61}, is in the common base configuration, with oscillations maintained by positive feedback from a tap on the collector load to the emitter. This design has been developed to give great stability of frequency for changes in supply voltage, and over the supply voltage range 15 to 25V the frequency changes less than ±0·2 per cent. In practice the supply line always lies between 22 and 25V, and the stability of the oscillator depends on thermal drifts in L_{61} and in its associated tuning capacitors. The frequency is selected by the switch S_{62}, the associated capacitors having been chosen to give the desired output frequencies (48 to 52c/s and 58 to 62c/s in 1c/s steps). The eleventh position of the switch S_{62} introduces no additional capacitance, and thus any desired frequency can be set by connecting a suitable external capacitor across terminals provided on the front panel. The twelfth position of S_{62} places R_{64} across the capacitors, thus turning off the oscillator, and allowing the input from the external oscillator, connected to appropriate terminals on the front panel, to become effective. The output from the oscillator is squared in the common emitter stage VT_{62}, which drives the emitter-follower, VT_{63}. An output for driving an external frequency meter is available at this point. The bistable binary divider stages VT_{64} and VT_{65}, VT_{66} and VT_{67}, are conventional, and the outputs from the collectors of the second stage are buffered in the emitter-followers VT_{68} and VT_{69}, driving the grids of thyratrons V_4 and V_5, via the step-up transformers T_{61} and T_{63}. The invertor itself is of conventional design, with a commutating choke L_{62} and a commutating capacitor, C_{85}. Snubbing components, C_{86}, R_{89}; C_{87}, and R_{90} hold the commutation factor of the thyratrons within specification. The output voltage is monitored by the conventional bridge rectifier, MR_{73-76}, and the output current by the current transformer T_{64} and the associated rectifier MR_{68-71}. R_{93}, the d.c. monitoring resistor has to withstand a normal load of 36W, and peak fault loads of as high as 400W. It is therefore of special design, using 12 s.w.g. constantan wire on a 1·5in diameter ceramic former. The switch, S_{61}, is for servicing purposes only and is normally always closed.

Performance

The invertor has been tested with a variety of loads and the output waveforms obtained are shown in Figs. 5 and 6. With loads predominantly resistive or inductive satisfactory operation is obtained at all loadings, but with loads that do not balance over the full cycle of output, such as electronic equipment with a half-wave rectifier drawing a substantial current (e.g. a domestic television set), trouble with regulation and with sporadic tripping may occur. This trouble may be overcome by adding a parallel resistive load. The invertor has run for several hundred hours at various loadings, and apart from the arcing back, referred to above, with the older type of thyratron, no difficulties have been encountered. There is no indication of gas clean-up in the invertor thyratrons, and trippings average about one every ten hours with normal loads. It is believed that most of these accidental trippings can be attributed to stray spikes on the mains input.

Conclusions

A relatively simple 1kW invertor using inert gas filled thyratrons, and showing good stability of output voltage and frequency has been described. The use of semiconductor devices in the control circuits enables all mechanical contactors to be dispensed with, thus restricting the duration of any fault condition. The invertor has proved reliable in service.

Direct-Coupled Silicon Transistor Switching Circuits

By J. D. Martin*, M.Sc.(Eng.), A.M.I.E.E.

The use of direct-coupled silicon transistor switching circuits is well known, but there is very little design data available. A study has been made of the problem, in order to design reliable circuits rather than ones possessing maximum gain. The transistor collector saturation voltage is discussed, and recommendations are made which enable design parameters to be estimated. Experience with circuits designed in this way has confirmed the usefulness of the procedure.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

TRANSISTORS are admirably suited to use in switching circuits, and this application is one where the majority of transistors are used. Low speed switching circuits (< 1kc/s say) are relatively straightforward to design and may readily be made to work over a wide range of operating conditions. Two of the most important aspects of such circuits are these:

(a) The ability to work over a specified temperature range.

(b) The ability to operate with any transistor of a given type.

Neglecting switching times, the transistor part of the problem is to ensure that the transistor collector current achieves a good on/off ratio. Fig. 1(a) shows a conventional switching stage. When the transistor is 'on' or saturated, the collector current is approximately V_{cc}/R_c since V_{ce} becomes very small.

When the transistor is cut-off, there is a small collector current still flowing, which depends on the base-emitter voltage and the transistor junction temperature (Fig. 2). If the collector current is to be reduced to its minimum value I_{co}, then the base-emitter voltage must be made more positive than $-V_{b1}$. Conventional pnp switching circuits use a positive bias supply to achieve this effect (Fig. 1(a)). For silicon alloy transistors, the saturated collector voltage V_{cs} may be made less than V_{b1}, giving cut-off conditions without the use of a positive bias supply[1] (Fig. 1(b)). The circuit may be further simplified under certain circumstances to that in Fig. 1(c), which is normally called a direct-coupled switching circuit. The cut-off problem is essentially the same in both cases. There are two characteristic properties of silicon transistors which make this type of connexion practical.

(a) High base-emitter voltage.

(b) Low leakage current.

Although silicon transistors are often operated without a bias supply, particularly in integrated transistor logic modules[2], very little data is available to help make a well designed circuit using conventional transistors. The details presented below are the results of an extensive study of the problems involved in this technique, and provide a guide to the type of silicon alloy transistor which is most suitable for direct-coupling. A useful analytical solution to the problem has not yet been discovered, so the discussion is largely based on empirical observations.

Having decided on the initial approximations and the method of applying the various measurements, the present aim is always to produce a worst-case design. Most of the discussion is therefore based on this worst case, although the majority of circuits constructed from these considerations would be working under more favourable conditions. The object of the exercise is to produce reliable circuits which will operate for many years, rather than to produce a design for maximum possible gain etc.

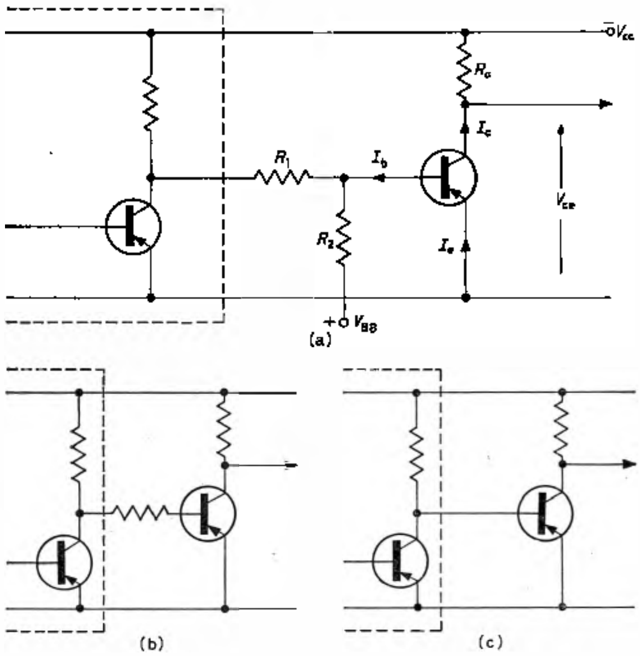


Fig. 1. Switching circuit variants

Consequently the approach becomes that of a cautious pessimist rather than that of a precise designer who is anxious to extract the last ounce of merit from each transistor.

Factors Involved in Circuit Design

Conditions for Cut-off

Collector leakage current I_{co} is very small at room temperature (~ 1nA), and is approached quite gradually by the collector current (Fig. 2). Consequently it is not convenient nor desirable to determine the base-emitter voltage which will reduce the collector current to I_{co}, and a voltage V_{bo} is obtained which reduces the collector current to an insignificant value. Fig. 3 shows V_{bo} as a function of temperature, as obtained from a large number of measurements on silicon alloy transistors of several types. This graph may be considered suitable for all transistors of the OC200, OC700, and 2S300 series, and for working collector currents greater than 1mA. A realistic margin of safety is included which helps to improve the circuit reaction to interference pulses. The requirement for satisfac-

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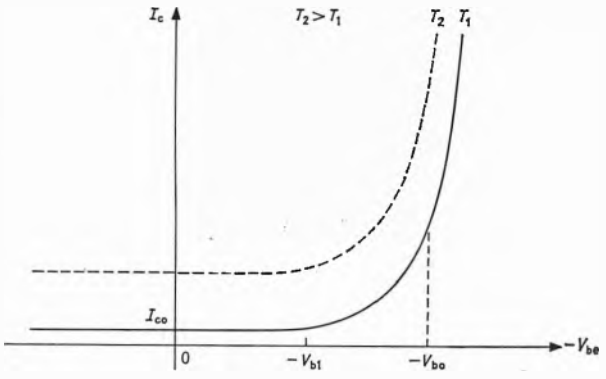


Fig. 2. Base voltage/collector current transfer characteristic

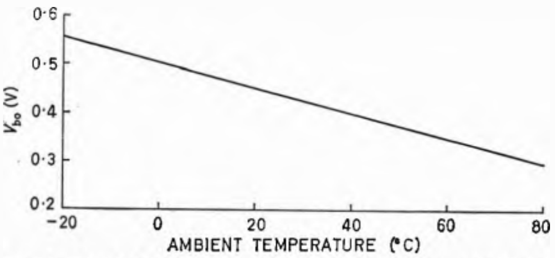


Fig. 3. Threshold base voltage V_{bo}, against ambient temperature

tory cut-off is that the collector voltage of the saturated driving transistor must be less than V_{bo} over the operational temperature range.

Collector saturation voltage V_{cs} is a function of collector and base currents for a given transistor. V_{cs} approaches a minimum value as the base current is increased, but this is outside the useful range of most alloy transistors since the base current is then of the same order as the collector current. The collector junction of the transistor includes a series resistance, which causes the saturation voltage V_{cs} to rise as the collector current I_c is increased. This effect is discussed in the Appendix. Fig. 4 compares four typical transistors and shows that because of the differences in collector series resistance, all types of silicon transistor are not equally suited to direct-coupling. The immediate effect of a large series resistance is to restrict the collector current range for which direct-coupling is possible. For instance, if $V_{bo} = 0{\cdot}2$V, then from Fig. 4, collector currents must be limited to 10mA for the OC700, and 16mA for the OC702.

When specifying a transistor for use in a switching circuit, the current gain is the most important single factor since it determines the base and collector resistor values. There are three current gains of importance in this instance.

- β is the normal large signal current gain for a specified collector current.
- β_D is the direct-coupled current gain; being the ratio of collector to base currents which gives conditions suitable for direct-coupling. Specified for a certain collector current and ambient temperature, thus when $I_c/I_b = \beta_D$, $V_{cs} = V_{bo}$ for that temperature.
- β_c is the circuit gain, being the ratio of collector and base currents in a physical circuit. These currents are determined by the collector and base resistors, and are basically independent of the transistor.

In general: $\beta_c \leqslant \beta_D \leqslant \beta$. The significance of these different current gains will become apparent as the discussion proceeds. β_D is a function of the normal current gain β and the collector saturation resistance. Fig. 5 shows the variation of β_D with collector current for a number of transistor types, at two different temperatures.

The current range restriction imposed by the collector series resistance is now clearly seen. The normal current gain β is increased at the higher temperature, but the direct-coupled gain β_D has decreased for the high collector currents; since the p.d. across the collector series resistance has become more significant compared with V_{bo}, which falls with increasing temperature.

Should a circuit design now be required involving one of the transistors whose characteristics are given in Fig. 5, then the chosen circuit gain β_c must be less than the minimum value of β_D over the specified temperature range, at the specified collector current. Reliable circuit design requires that the final circuit must work with any transistor of the chosen type. Consequently curves such as those in Fig. 5 should be available for the worst transistor of the type.

Transistor Selection

There are two general factors which will make a transistor suitable for direct-coupled operation.

(a) Reasonable current gain β at the working collector current.

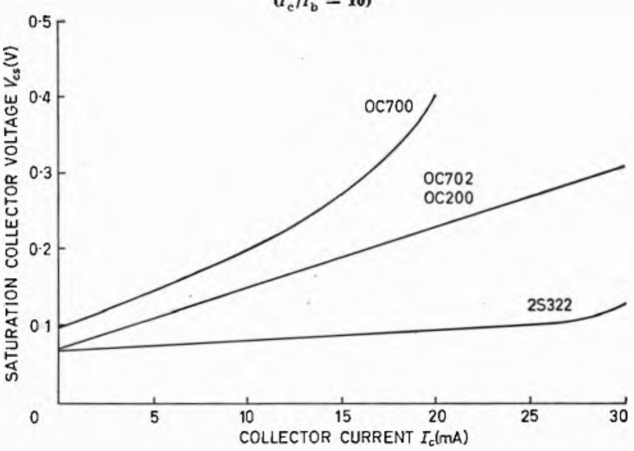
(b) Low collector saturation resistance.

It is necessary to compare a number of commercial types in order to decide which ones are suitable for direct-coupling under given conditions. The end result of the assessment of a transistor type, is the minimum value of β_D to be expected from any transistor of the type at specified collector currents and ambient temperatures.

The saturation resistance is not normally given explicitly by the manufacturer, but an average value can be determined from the published collector characteristics, by constructing a graph of V_{cs} versus I_c with β_c/β constant, after the style of equation (3) in the Appendix. A maximum value may be estimated from the maximum saturation voltage (which is normally given), provided that the inverse current gain α_I is known. (Equation (1), Appendix). Since the value of α_I cannot be determined with any accuracy, and the quoted saturation voltage is a limit rather than a parameter, the estimate of saturation resistance is unreliable.

In order to obtain a reliable estimate, it appears that an empirical method must be used. A representative batch of transistors of a given type may be tested for β_D, and also for β under the conditions quoted by the manufacturer. This latter measurement then provides a control comparison which may be used to adjust the minimum values of the measured β_D, to take first order account of the spread in characteristics of the total population of this

Fig. 4. Collector saturation voltages of four typical silicon alloy transistors ($I_c/I_b = 10$)



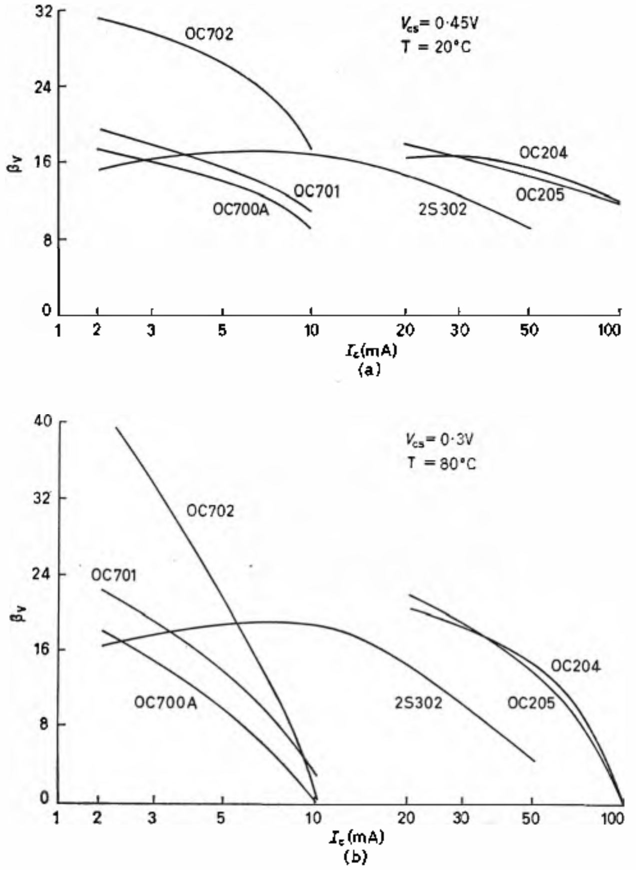


Fig. 5. Measured variation of β_D with collector current and temperature

N.B. Each curve represents the worst case from a batch of 25 transistors of the type: (a) 20°C ambient; (b) 80°C ambient

transistor type. For instance, if the measurement of β shows a reasonably normal distribution with the minimum measured value 20 per cent higher than the manufacturer's quoted minimum, the implication is that the measured minimum value of β_D will be also about 20 per cent higher than the absolute minimum to be expected. This argument is empirical rather than precise, since batches of purchased components rarely follow a normal distribution, but in any case a precise answer is only of little more practical use than a result obtained as indicated. Fig. 5 represents the minimum gains obtained experimentally from batches of 25 samples of each of the transistor types.

In practice it is not necessary to obtain the values of β_D over a wide temperature range. If the base and collector currents are held constant, collector saturation voltage V_{cs} increases very little as the ambient temperature is increased from 20°C ($V_{cs} \propto$ absolute temperature). Consequently, if β_D is measured at 20°C, but with $V_{cs} = V_{bo}$ for the maximum temperature required, then this value of β_D will be satisfactory for all temperatures in this range.

Conditions at the lowest temperature are not so readily determined without actually reproducing them, since the reduction of normal current gain β with temperature tends to reduce the degree of saturation and hence raise V_{cs}. However the change in gain is only about 10 per cent for a 40°C drop in temperature from 20°C, so that provided a suitable factor of safety is taken, it is almost certain that even the lowest gain transistor will maintain a satisfactory saturation voltage at the lowest temperature. The measurement at 20°C can therefore serve for the whole temperature range.

Circuit Considerations

The measurements discussed in the foregoing section enable a particular transistor type to be selected. The circuit gain β_o which determines the actual circuit resistor values, is given by: (Min.β_D)/(Factor-of-safety). This factor-of-safety can be chosen freely, and covers the following uncertainties.

(a) In the operating conditions estimated as above.
(b) Induced interference signals.
(c) Change in transistor gain with life.
(d) Component changes with life.
(e) Misuse of the circuits.

Uncertainties in operating conditions (a) are clearly feasible due to the empirical nature of the representative measurements made above. Some account of interference has already been taken in determining suitable values for V_{bo}. However this factor refers to the driving on of a cut-off transistor due to an interference signal exceeding the base threshold level; whereas the gain factor-of-safety prevents an interference signal cutting an on-transistor off due to the base current being reduced. While these two effects are slightly related, it seems necessary to make two independent corrections owing to the non-linear relationship between them.

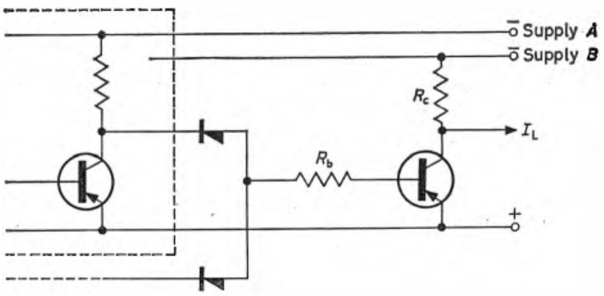
While it is not possible to take full acount of circuit misuse (e), it is a factor to be considered if at all possible. These circuits normally form elements of a complex connected unit, and unintentional loading errors as well as unforeseen ambient conditions can occur. If each individual circuit is able to cope with conditions somewhat worse than designed for, this can save much time and effort when these extraordinary conditions occur, particularly in a large system.

In the author's experience a gain factor-of-safety of 2 has been taken. Although this drastically reduces the circuit gain, it does provide good isolation from the effects mentioned above. This loss of gain is not disastrous for switching circuits, although it may mean that transistor-diode logic will be necessary. A smaller factor-of-safety may be satisfactory for laboratory and manufacturing environments, but this margin of safety could become dangerously low over a period of 10 or 20 years, when interference and component drift are considered.

Leakage Current

In direct-coupled circuits, the cut-off collector current has been reduced to a suitably low value by keeping the saturation collector voltage of the preceding transistor below V_{bo}. In cases where series diodes are used in the base circuit (Fig. 6), the base circuit is effectively open-circuited at cut-off. Consequently the collector leakage current is approximately $I_{co}' = \beta I_{co}$, which may be large enough at high temperatures to become a nuisance. Since I_{co}' is not normally quoted by the manufacture for silicon alloy transistors, its value must be estimated from the current gain β at a low current level, and I_{co}. Tests

Fig. 6. Switching stage



indicate that there is no definite correlation between β and I_{co}, so that the maximum possible value of each should be taken. A large factor-of-safety must be included since leakage current may increase greatly during equipment life. This leakage current will flow in the collector resistor R_c reducing the effective supply voltage by $I_{co}'R_c$. Perhaps a suitable criterion of satisfaction is that $I_{co}'R_c$ should represent less than 10 per cent of the supply voltage.

Should it be necessary to reduce the leakage current below I_{co}', then a resistor R may be connected between base and emitter. Since I_{co}' increases with temperature it is now only necessary to ensure that $I_{co}'R$ is less than V_{bo} (Fig. 3) at the highest ambient temperature considered. Under these conditions, the leakage current is reduced to the order of I_{co}.

Marginal Checking

It has been the practice for many years with switching circuit systems to apply an overall test by increasing the bias supply. The base current at each stage is reduced, and hence each individual transistor is checked for mal-operation under conditions more arduous than normal. This same method of testing is not possible with direct-coupled stages. However, a similar effect may be obtained by taking the collectors of alternate stages to different supplies (Fig. 6). Differential variation of these two supplies now reduces the base current in alternate stages, and allows the check to be applied. In a complex system, it is not always possible for every transistor to be connected in this way. In practice, however, the vast majority may be connected thus, and the marginal check has proved to be useful in fault-finding at development and manufacturing stages.

Example

In order to illustrate the use of the different factors, an example will be introduced. Suppose that the 2S302 transistor is being studied, and that the curves in Fig. 5 represent the worst transistor of this type. It is required to design two switching stages, with collector currents of 2mA and 30mA respectively, to operate over the temperature range 20 to 80°C. Table 1 summarizes the results from the graph. The lowest value of β_D must be taken in each case, and a suitable factor-of-safety introduced to produce the circuit gain β_c.

Taking a factor-of-safety of 2, the circuit gains β_c are 5·25 and 7·5 for collector currents of 30mA and 2mA respectively. Thus base currents of 5·7mA and 0·27mA are required. The resistor values follow from these currents, taking the supply, input, and base-emitter voltages into account. Notice that for 2mA collector current, the lowest gain occurs at 20°C, whereas it occurs at 80°C for 30mA collector current. This is due to the influence of the collector saturation resistance.

The maximum collector leakage current at 80°C is found from the manufacturer's data to be 1·8μA. The maximum current gain β at the corresponding level of cut-off collector current is about 36. Consequently a collector leakage current of up to 65μA may be expected in a circuit as in Fig. 6. Applying a factor-of-safety of 4 since leakage current tends to vary logarithmically, means that the circuit must work satisfactorily with a cut-off collector current of 260μA. Although there is no problem with the 30mA circuit, this leakage current is greater than 10 per cent of the on collector current in the 2mA circuit. Consequently the maximum resistance which can be allowed in the base-emitter circuit is 0·3/1·8 = 0·17MΩ.

Readings from Fig. 5 for 2S302

TABLE 1

I_c (mA)	T (°C)	β_D
30	20	13
30	80	10·5
2	20	15
2	80	17

It should be noted that with a gain factor-of-safety of 2, the differential movement of the marginal checking supplies, can be up to nearly 50 per cent of the supply itself. This gives an indication of the ruggedness of such a circuit in adverse conditions.

Several hundred similar circuits have been type tested and operated correctly over the required temperature range, with the supply voltage reduced by as much as 70 per cent of nominal.

Many thousands of similar circuits have been constructed using these principles, and no difficulties have arisen due to mal-operation. The conservative approach to design has been justified by the absence of circuit-mal-function failures in a large number of complex systems.

Conclusions

A reasonable basis has been presented, for operating silicon transistor switching circuits without a bias supply. The collector saturation resistance tends to limit the useful current range of such a connexion, but it cannot be determined with any accuracy. An empirical method is suggested for selecting transistor types and assessing their gain capabilities. Leakage currents may become important if a transistor base circuit is open-circuited to produce cut-off, but the leakage current may be greatly reduced by connecting a resistor between base and emitter. Marginal checking is shown to be possible with these circuits.

The principles stated here are the result of a great deal of practical work, and have proved their worth in the fault-free operation of circuits involving many thousands of transistors.

Acknowledgments

The author would like to thank the Westinghouse Brake & Signal Co. Ltd for permission to publish these details, and for initiating the project which led to this study.

Appendix

Collector Saturation Voltage V_{cs}

Ebers & Moll[3] have calculated the collector voltage in saturation neglecting resistive components. Using their equation, and writing:

$$\beta_c = I_c/I_B,\ \beta = \frac{\alpha_N}{1-\alpha_N}$$

The internal voltage

$$V_{ce}' = kT/q\ \ln\left\{\frac{1+(1-\alpha_I)\,\beta_c}{\alpha_I\,(1-(\beta_c/\beta))}\right\} \quad \ldots\ldots\ldots\ldots\ldots (1)$$

α_I = inverse current gain
k = Boltzmann constant
T = absolute temperature (°K)
q = electronic charge

Suppose now, that series resistances at collector and emitter junctions, r_{cs} and r_{es}, are considered. These will be due to body resistance of the contact materials, and imperfections of the junctions.

Then external voltage

$$V_{cs} = V_{ce}' + I_e \,.\, r_{es} + I_c r_{cs} \quad \ldots\ldots\ldots (2)$$

Substituting $I_e = I_c\,(1 + (1/\beta_c))$

$$V_{cs} = V_{ce}' + I_c\,(r_{cs} + r_{es}) \quad \ldots\ldots\ldots\ldots (3)$$

since $\beta_c \gg 1$ and $r_{es} < r_{cs}$

This equation is of the general form $V_{cs} = V_o + I_c R$ where V_o is approximately constant if β_c/β is constant.

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A Precision Three-Phase 400c/s Generator

By D. J. Miles*

A three phase supply which is stable in amplitude as well as frequency has been developed as a wheel power supply for inertial quality gyroscopes. Amplitude stability of ±0·01 per cent against normal load and mains supply variation is achieved.

(Voir page 776 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

TESTING of inertial quality gyroscopes has shown that the drift performance of a gyroscope can be significantly affected by amplitude variations in the power supply to the gyroscope wheel. A three phase 400c/s supply has accordingly been developed with an amplitude stability of ±0·01 per cent for use with loads in the 0 to 50W range. (A typical gyroscope wheel requires 5 to 10W.) The unit may be driven from a single phase 400c/s crystal controlled source, from which amplitude variations of ±5 per cent can be tolerated, but has it own internal *LC* oscillator for use when precise frequency control is not required.

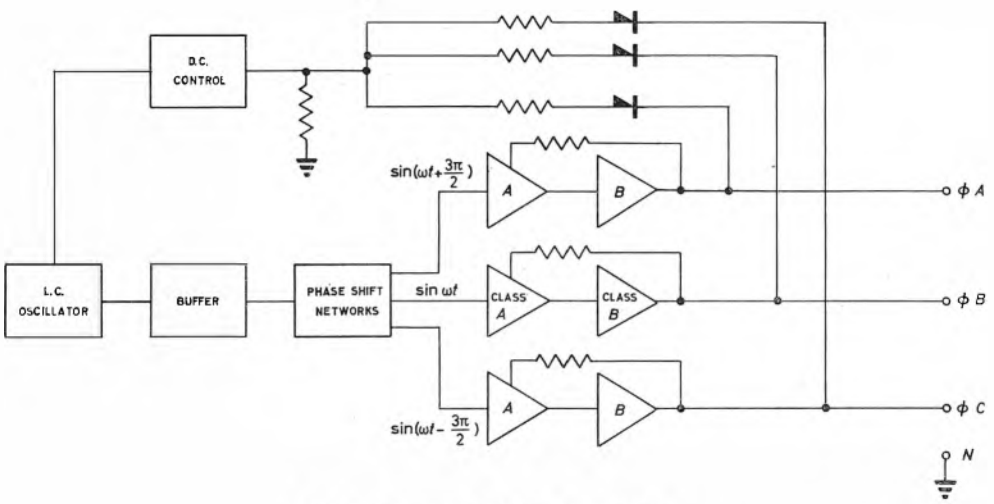


Fig. 1. Arrangement of three-phase supply

Principle of Operation

Fig. 1 is a block diagram of the unit. The output from the *LC* oscillator is buffered and fed to phase shift networks which produce two outputs phase shifted by +120° and −120°. These are passed, together with a third output at 0°, to three class-A voltage amplifiers followed by class-B power output stages. A clean and reasonably stable three phase output is then produced by overall a.c. feedback from the outputs of the class-B stages to the class-A amplifier inputs. However, the amount of a.c. feedback which can be applied is obviously limited (unless complicated shaping networks are employed) and, furthermore, the oscillator buffer and phase shift networks are outside this feedback loop. Consequently an overall control loop has been added in which the mean of the three output voltages is obtained and compared with a d.c. reference voltage. The difference between these voltages is amplified and used to control the h.t. supply to the oscillator (the a.c. output of which is proportional to the h.t. supply) in the sense required to bring the difference to zero. In this manner the output can be controlled to within the limits of accuracy of the reference voltage (for which a low temperature coefficient Zener diode is used) and the measuring resistors.

* *Royal Aircraft Establishment.*

Practical Networks

Oscillator, Buffer and Phase Shift Networks

The oscillator, shown in Fig. 2, is of conventional *LC* form which utilizes an adjustable inductor wound on a 'Vinkor' pot core for frequency trimming. The oscillator amplitude is practically equal to the applied h.t. voltage over the control range thus providing linear control of the output by h.t. variations. When the oscillator is driven from an external source the internal feedback is disconnected to avoid 'locking' problems, VT_1 and the *LC* network simply forming a tuned, h.t. limited, amplifier.

Phase shift networks R_6, C_3 and R_8, C_4 utilize the split supply obtained from centre-tapped transformer windings to provide adjustable phase voltages whose amplitudes are independent of the phase displacement. Setting up procedure is thus simplified since it is not necessary to reset the output amplitudes after phase adjustment.

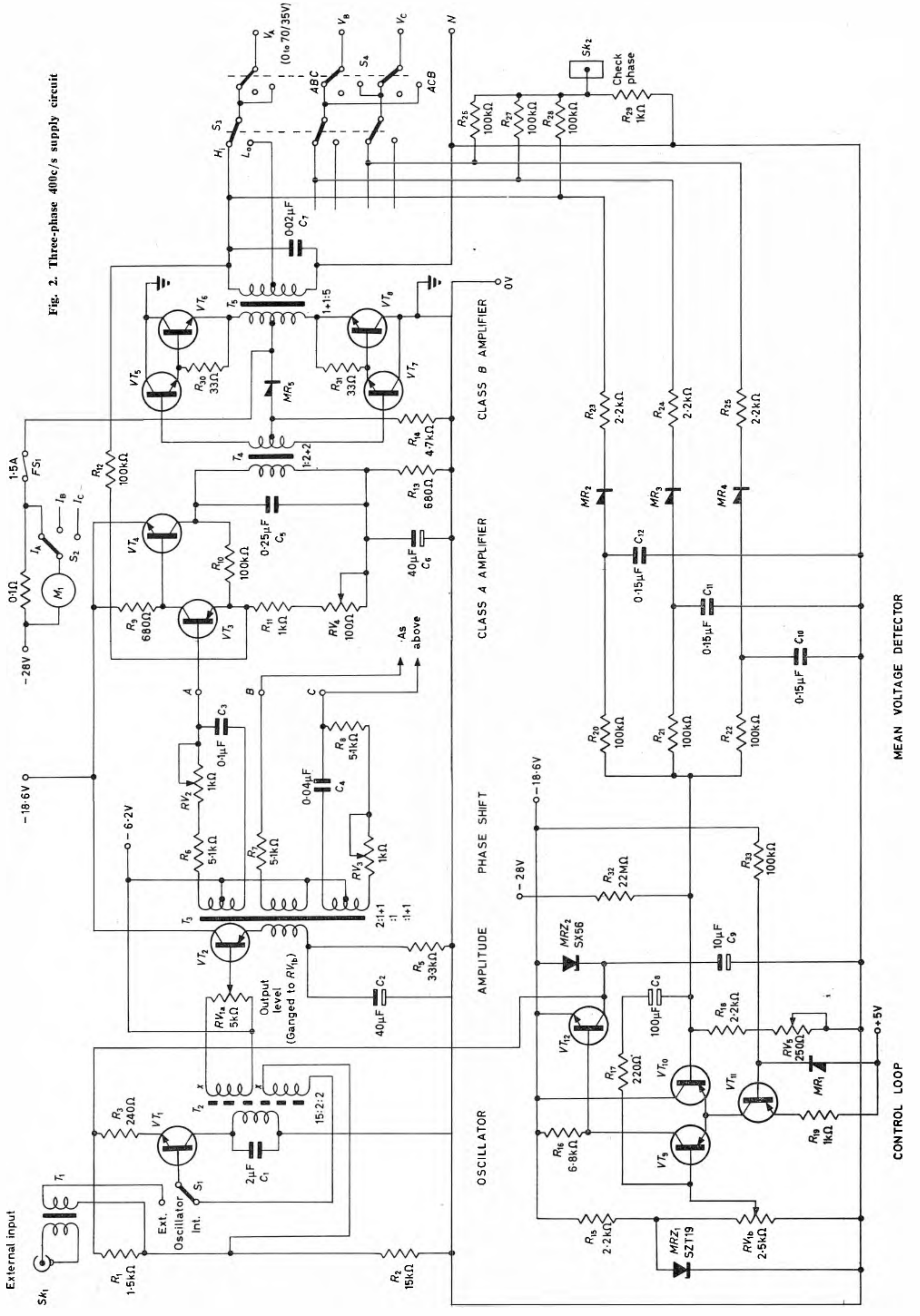
Class-A Amplifier

The class-A amplifier formed by VT_3 and VT_4 (ZT84) has feedback over two stages (R_{10}, R_{11}) defining a voltage gain of about 100 at 400c/s. This high gain is achieved without sacrifice of operating point stability since the transformer primary forms a direct feedback path to make the d.c. gain unity. Only one decoupling capacitor is used (C_6), C_5 being added to tune transformer T_4 and to reduce the loop gain (approximately five at 400c/s) at high frequencies.

Class-B Amplifier

The class-B output stages use compounded silicon transistors (ZT92, 2SO24) in a common collector mode. This configuration gives a low output impedance so that overall feedback may safely be taken from the output since the loop gain is then substantially independent of the load. The common collector configuration also gives low crossover distortion, which is further reduced by overall feedback, without the need for a high standing current. Some forward bias is provided by the silicon diode MR_5 which

Fig. 2. Three-phase 400c/s supply circuit



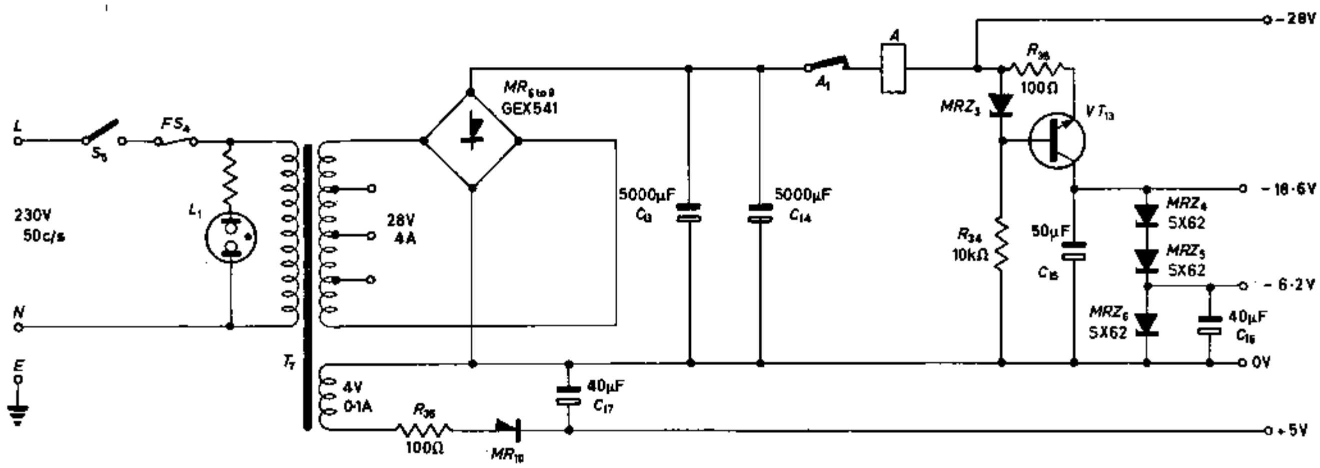


Fig. 3. Power supply circuit

is mounted on the same heat sink as the output transistors since tracking with the operating temperature is obviously more important than tracking with ambient temperature variations. The class-B stage is operated with collectors earthed to the chassis to obtain maximum heat transfer.

Mean Voltage Detector and Control Loop

Each of the three phase-to-neutral voltages are first rectified (MR_{2-4}) and then smoothed (C_{10-12}). Mixing resistors R_{20-22} then pass currents into R_{18} to produce a direct voltage, at the base of VT_{10}, which is proportional to the mean of these three phase-to-neutral voltages.

Transistors VT_9, VT_{10} and VT_{12} form a simple d.c. differential amplifier, the output of which is taken from the collector of VT_{12} to provide the oscillator h.t. supply. With the loop closed the oscillator h.t. supply is changed by the amplifier until the mean three phase output is such that the voltage at the base of VT_{10} is equal to the voltage level selected by the potential divider RV_{1b}. The amplifier is designed to operate over the small range (± 10 per cent output variation) necessary to contend with the effect of load and supply changes; selection of the required three phase output voltage is made by RV_{1a} which is ganged to RV_{1b}. The oscillator therefore always operates at a reasonable level (its h.t. is prevented from falling below 13V by MRZ_2) and the operating currents of the d.c. amplifier remain essentially the same for three phase outputs from 7V r.m.s. to 70V r.m.s.

The control loop has both proportional and integral gain, the proportional value being determined by R_{17} and the integral time-constant by C_8. The stability with the values indicated in Fig. 2 is adequate—a step load of 20W producing a 1 per cent transient change in output level with recovery to the 0·1 per cent region within four or five cycles of the oscillator. The loop gain is high, full load regulation of 2 per cent open loop being reduced to 0·02 per cent closed loop. The effect of mains supply variations of ± 5 per cent is virtually eliminated by feeding into the control loop a proportion of the -28V unregulated line via R_{32}.

Power Supply

The power supply is shown in Fig. 3. Full wave rectification followed by capacitance smoothing (C_{13}, C_{14}) provides the -28V unregulated rail; the $-18{\cdot}6$V and $-6{\cdot}2$V lines are derived by operating VT_{13} as a constant current source to reference diodes MRZ_{4-6}. The supply is protected against excessive loads by latching relay A which breaks at a current of 4A.

A $+5$V rail, which is only used by the control loop, is made necessary by the requirement that the output voltage should be continuously variable down to zero.

Metering and Setting-up Facilities

The unit contains a voltmeter (not shown in the circuit diagram) which may be switched to indicate any of the six phase-to-phase and phase-to-neutral voltages. The output current is shown on moving-coil meter M_1 which simply indicates the current flowing into the associated power amplifier. Good agreement is obtained between this reading and values measured at the output while at the same time one less resistance has been included in the output circuit. For the same reason the fuses are also placed in the amplifier h.t. line rather than in the output leads.

Setting up the instrument consists simply of:

(1) Adjusting all three output voltages to be equal by means of the two gain potentiometers in phase A and phase C class-A amplifiers.

(2) Setting frequency to 400c/s by adjusting the inductance of T_2.

(3) Adjusting the phases to 120° displacement by nulling the voltage at SK_2 with phase shift potentiometers RV_2 and RV_3. (Since the voltages have been equalized and the mixing resistors are equal this null can only be obtained when the three voltages are displaced by 120°.)

(4) Providing maximum control tolerance by adjusting RV_5 so that the voltage across VT_{12} is about $2\frac{1}{2}$V.

Characteristics

Steady State Performance

Voltage range:	High 0 to 120V r.m.s.	line to line
	Low 0 to 60V r.m.s.	
Current:	High 0·3A r.m.s. maximum	
	Low 0·6A r.m.s. maximum	
Frequency:	400c/s	
	Stability ± 1c/s on internal drive	

Harmonic distortion: 2 per cent

Amplitude stability and regulation: $\pm 0{\cdot}01$ per cent.

The Effect of Supply and Load Variations

The effect of mains supply, oscillator drive voltage and load power changes on the output amplitude are best seen by examining Fig. 4. This is a nine-hour recording of the difference between the rectified mean three phase output and a precision reference voltage, recorded on a potentiometric chart recorder. Loads of 0, 20 and 50W were applied under various conditions of oscillator drive (external and internal) and of mains supply voltage. The mains supply voltages noted were, in fact, the nominal value set by a variable transformer and were subject throughout to the normal mains supply fluctuation of up to ±5 per cent.

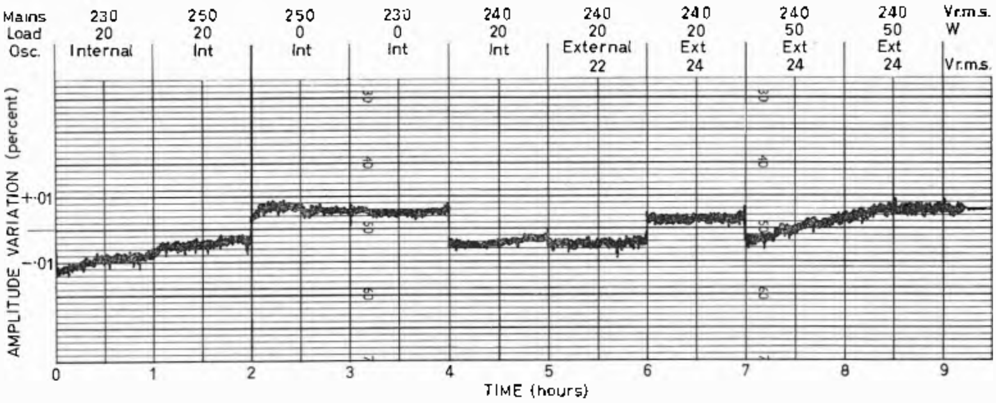


Fig. 4. Percentage variation of output amplitude as a function of time, mains supply voltage, load and drive amplitude

It can be seen from the recording that the output is within about ±0·01 per cent of the mean under all the imposed conditions and that, at the light loading associated with a gyroscope wheel, the stability approximates to ±0·001 per cent. This test was performed with an output of about 115V; a repeat with the output level halved showed the same stability.

Transient Response

Step load changes of about 20W were applied to the output and the amplitude photographically recorded. Transients of about 1 per cent amplitude were produced, the unit recovering with one overshoot, to within about 0·1 per cent in four or five cycles of the oscillator i.e. about 12msec.

Earthing Requirements

As shown in Fig. 2 the unit gives a three phase four wire output with an earthed neutral point. When a three wire, earthed phase, system is required either an isolating winding must be added to T_5 or else small isolating transformers added to provide the feedback. The latter arrangement is essential if the very low output impedance of the unit is required since, in the former arrangement, T_5 secondary resistance is outside the feedback loop.

Conclusions

A three phase supply has been developed which, by the inclusion of an overall control loop comparing the output voltage against a reference voltage obtained from a Zener diode, maintains the three phase output voltage constant within ±0·01 per cent under all normal operating conditions. The design is such that there are no 'select on test' components and only the essential minimum of six adjustments (one frequency, two phase and three amplitude) are necessary.

Acknowledgments

Acknowledgment is made to the Ministry of Aviation for permission to publish this article. The assistance of section members Mr. R. B. Syms and Mr. T. D. C. Carr and the technical advice of Mr. A. Cowie is also acknowledged.

Colour Television for Pilot Training

The BOAC VC10 flight simulators, at the airline's training centre adjacent to London Airport, are now operating with a new three-dimensional closed circuit colour television system that enables pilots to practice take-offs, approaches and landings under any desired weather conditions by day or by night—without leaving the ground.

Simulators have made a valuable contribution to BOAC flying staff training since the days of the Boeing Stratocruiser, but their usefulness has been limited by the fact that all simulator flying has been "blind"—carried out entirely on instruments with the flight deck windows blacked out. The latest aid to realism represents a big improvement in training facilities and it is expected to result in a significant reduction in conversion and check flying training with a consequent saving in cost.

A further benefit from a reduction in the time required for flying training is the increased availability of aircraft for revenue-earning flying.

Designed and manufactured by the Flight Simulator Division of Redifon Ltd, the visual system enables the pilot to look ahead through the windscreen and see the runway and the countryside around the airport in full colour as he would if flying a real aircraft.

This is achieved by using a television camera with a special lens system that moves over a large model landscape representing an area measuring 9 miles by 2½ miles. The camera lens system faithfully responds to the movements of the aircraft controls on the simulator flight-deck and the picture it sees is transmitted to a colour projector mounted on the roof of the simulator cabin. The picture is beamed on to a screen a few feet ahead of the cabin, and the flight-deck and screen together move in harmony with the movements of the controls by the pilot.

The large vertical model landscape, 48ft long and 13ft high, represents a rural scene with fields, hedges, woods, roads and small towns and villages. It is powerfully lighted by banks of tungsten lamps using some 180kW of electricity. The television camera is mounted on a gantry that travels on a railway track between the face of the model and the banks of lighting reflectors.

The system enables pilots to make a visual approach to the runway, co-ordinated with the simulator instrument landing system equipment, from a distance of five miles and a height of 1 500ft. Visibility can be varied from 4 miles to down to 100 yards and cloud and fog effects may be simulated over the full height range from 1 500ft to ground level. The model of the airfield is also equipped with runway and approach light and angle of approach indicators lighting, so that night-time landings and take-offs can be simulated.

A black and white monitor television screen is installed at the instructor's control panel so that he can view the picture being seen by the pilot, and can introduce drift to simulate cross-wind landings as well as determining the degree of visibility and the other conditions under which a landing is being made.

Amplifier Design with a Digital Computer

By J. R. Abrahams*

This article shows how a small digital computer may be used to perform design calculations for electronic circuits. The design of a two-stage tuned amplifier, employing transistors, is used as a numerical example, with some simplification of the parameters. The possibility of doing comparative cost calculations, and thus arriving at the economically optimum design, is also discussed.

(Voir page 776 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

THE growing availability of the digital computer will give many engineers the opportunity to use automatic computation as an integral part of circuit and system design. In this work it is desirable that the engineer should be capable of writing his own programs, at least in general terms. The use of flow graphs, as the intermediate step between a physical system and its simulation in the computer, together with a problem-orientated language, such as Fortran or Algol, has proved to be very helpful for this purpose.

Flow Graph Methods

The theory of signal flow graphs (or, more generally, flow graphs) has been considered by several authors[1,2] and is not repeated in detail here. However, it may be helpful to state a few of the basic ideas, in order that the computer programs may be fully understood. A flow graph is a pictorial method of representing a set of algebraic equations, and consists of nodes linked by unidirectional branches. Each variable in the set of equations is represented by a node and the coefficient relating any two variables is included in the flow graph as the transmission of the branch linking the two appropriate nodes. Fig. 1 shows the flow graph for the two equations:

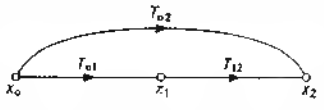


Fig. 1. Simple flow graph

$$x_1 = T_{01}x_0$$

and

$$x_2 = T_{02}x_0 + T_{12}x_1$$

The usual problem is to find the resultant transmission through a flow graph, from source to sink. For a simple graph this may be done manually, often with the aid of Mason's rule[3]. For larger graphs, particularly where the branch transmissions are complex and variable, it is almost essential to use a computer, to reduce a flow graph having many branches to a graph having only one branch, linking the two desired nodes. Three simple flow graph rules are used in this reduction process, and these may be stated as follows:

(a) The transmissions of branches in parallel are added.

(b) The transmissions of branches in series are multiplied together.

(c) The effect of a loop through a node is to divide all other transmissions through that node by a factor of (1—transmission round the closed loop). This is the familiar feedback equation.

These rules are illustrated by the three flow graphs of Fig. 2, which also includes the corresponding algebraic equations.

A simple Fortran program which may be used to reduce a large flow graph to a two-node, one-branch, graph is given as part (a) of the Appendix. In this program the branch transmission from node I to node J is given as $T(I,J)$ and the program can only handle graphs having real transmissions. The maximum size of flow graph which can be reduced by this program is limited by the storage capacity of the digital computer. All the programs in this article were run on an I.B.M. 1620 having a high-speed

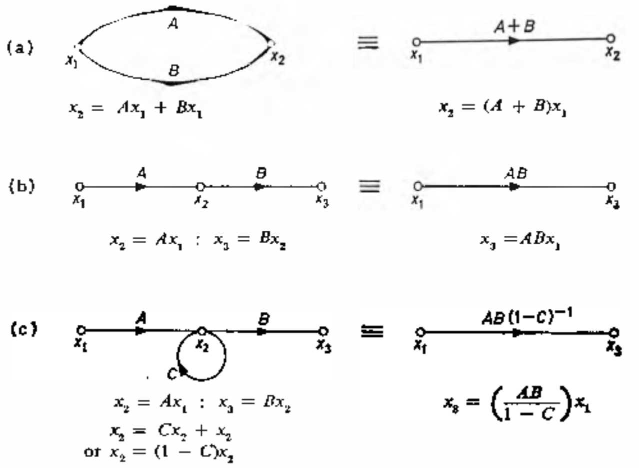


Fig. 2. Basic rules for flow graphs

store for 20 000 decimal digits (i.e. 2 000 ten-digit numbers in Fortran), which allows for a flow graph of up to 30 nodes. The majority of electrical systems have a flow graph containing only a small proportion of possible branches linking the nodes. This means that most of the branch transmissions in the program are, in fact, zero. As it stands program (a) is very inefficient for large graphs, in that there is no test to check for zero transmissions and therefore most of the multiplications would produce a zero result.

The detailed step-by-step procedure to reduce a flow graph by node elimination has been given previously[4]. For a flow graph, where NN is the number of nodes, program (a) starts by eliminating node 2 and proceeds until node $(NN - 1)$ has been eliminated. The numbering of nodes within a flow graph may be done in any order, provided that the source of interest is node 1 and the sink is given the highest node number.

The operation of program (a) may be tested with the two flow graphs of Fig. 3. The source-to-sink transmissions for these two graphs may be calculated mentally, and the results from the computer are given in the Appendix.

One disadvantage of this simple reduction program is that all the branch transmissions must be read into the computer, in the form of a transmission matrix. In a typical case this would mean that another program is needed to produce a long data tape, or deck of punched

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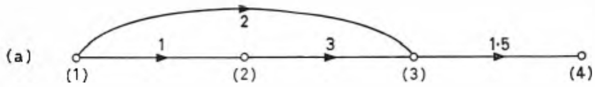
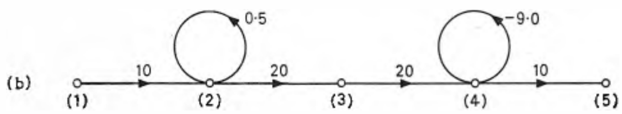


$NN = 4$

$T_{14} = (1 \times 3 + 2)\ 1{\cdot}5 = 7{\cdot}5$



$NN = 5$

$$T_{15} = \frac{10 \times 20 \times 20 \times 10}{(1 - 0{\cdot}5) \times (1 + 9)} = 8\ 000$$

Fig. 3. Two test flow graphs, for computer program

cards. It is therefore usual to form the branch transmissions within part of the main program, and this is done in the following example.

Amplifier Circuit Example

The tuned amplifier circuit of Fig. 4 is used as the basis of a numerical example, to illustrate the flow graph and programming procedures.

This two-stage circuit, with the input transistor in the common emitter mode and the output transistor in common base, has been successfully used in i.f. amplifiers at 10·7Mc/s and 60Mc/s. The transistor pair gives the characteristics of a cascode amplifier, with medium input resistance, high output resistance and small voltage feedback. The analysis of this circuit and the related transistor parameters have been considered fully by Coverley[3], on a flow graph basis. The basic inter-connexions for the two transistors of this amplifier are shown in Fig. 5, which also gives the voltages and currents of interest.

The flow graph for this amplifier may be drawn in terms of the hybrid parameters, and this is done in Fig. 6. In this flow graph the input voltage (V_i) is shown as derived from a generator with a source admittance of Y_G and the effective resistance of the tuned load is taken as R_L, with an output voltage V_o.

This flow graph is constructed by first drawing the h-type flow graphs for each transistor, the left-hand one being for the common-emitter mode. These rectangular flow graphs are linked by unity transmission branches of the appropriate sign, by inspection from Fig. 5. Clearly V_{12} and V_{21} are equal and $i_{12} = -i_{21}$.

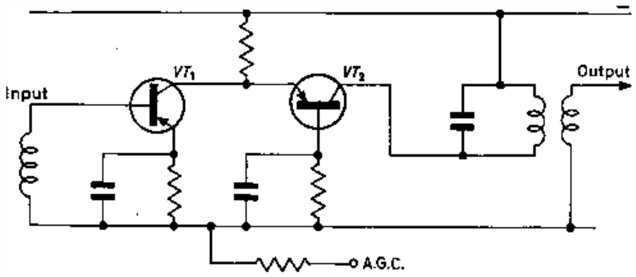
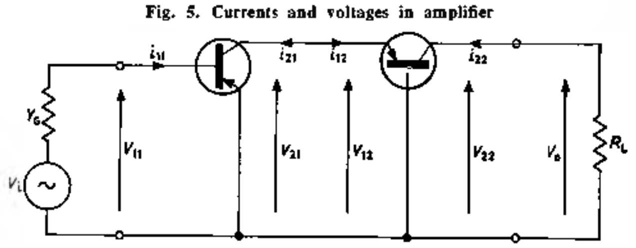


Fig. 4. Two-stage transistor amplifier

Fig. 5. Currents and voltages in amplifier



The branch transmissions linking the source (V_i) and sink (V_o) to the flow graphs are derived from the equations that:

$$i_{11} = Y_G (V_i - V_{11}),$$
$$V_{22} = -i_{22}R_L,$$
$$V_o = V_{22}.$$

To produce a meaningful example it is now necessary to assume some numerical values for the parameters in Fig. 6. For this purpose it is considered that the direct emitter current is the same for both transistors, that the two transistors are of the same type and that the current gain, in common-emitter varies in the manner shown by Table 1. The results of Table 1 were obtained, at the frequency of interest, by a simple method given by Pridham[5].

At low frequencies the hybrid parameter h_{11} is related to the direct emitter current and a relationship of this form is used here, where the symbol $C(N)$ indicates emitter current and $H(N)$ the corresponding common-emitter current gain. There is a complicated relationship between the h_{21}' and h_{21}, but in this case it is taken that $h_{11}' = 20\,h_{11}$. The fixed numerical values employed in this

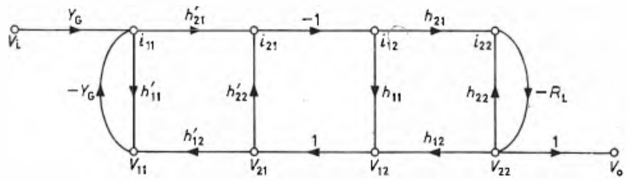


Fig. 6. Flow graph for two-stage amplifier

example are as follows:

$h_{12}' = (7 + 3j)\ 10^{-3}$, $h_{12} = (5 + 2j)\ 10^{-3}$

$h_{22}' = (4 + 1j)\ 10^{-3}\text{mho}$, $h_{22} = (1 + 15j)\ 10^{-4}\text{mho}$

$Y_G = (25 + 5j)\ 10^{-4}\text{mho}$, $R_L = 500\Omega$

The expressions for h_{21}' and h_{21} take account of the phase angle involved (which is around 20° for the common-emitter mode) and hence the variable parameters in the flow graph of Fig. 6 are:

$h_{11}' = (500+120j)\ (1/C(N))\Omega$, $h_{11} = (25+6j)\ (1/C(N))\Omega$

$h_{21}' = -(0{\cdot}94+0{\cdot}35j)\ H(N)$, $h_{21} = -(1+0{\cdot}02j)H(N)/[1 + H(N)]$

Computer Program

The complete computer program consists of several clearly defined parts, which are given as (b), (c) and (d) in the Appendix. This arrangement is desirable to save effort in preparing the source tape (or cards). The first two procedures are quite general and can be used for any flow graph having complex branch transmissions. The transmission of a branch from node I to node J is given

TABLE 1
Emitter Current and Gain Values

MEASUREMENT STEP NUMBER N	EMITTER CURRENT $C(N)$ mA	CURRENT GAIN $H(N)$
1	0·1	$-26\angle 17{\cdot}5°$
2	0·3	$-31\angle 18°$
3	0·5	$-37\angle 19°$
4	0·7	$-42\angle 19{\cdot}5°$
5	1·0	$-45\angle 20°$
6	2·0	$-47\angle 20°$
7	3·0	$-41\angle 20{\cdot}5°$
8	5·0	$-36\angle 20{\cdot}5°$
9	7·0	$-32\angle 21°$
10	10·0	$-25\angle 21°$

by its reference ('real') and quadrate ('imaginary') parts $R(I, J)$ and $Q(I, J)$. Procedure 60 does some of the basic complex arithmetic which is required and also computes the source-to-sink transmission through the graph in amplitude (AN) and phase angle ($PH\,1$). It also computes the inverse of the through transmission (BN) and the associated angle ($PH\,2$). Procedure 70 performs the operations to eliminate any node L, where the value of L goes in turn from 2 to the penultimate node number. For this example a simplified version of procedure 70 was used, including two *IF* statements to check for zero transmissions. In fact only the reference part of each transmission is checked, knowing that in the frequency range under consideration no branch can have a transmission which is wholly quadrate. For this short procedure to be completely general it would be necessary to check for all possible zero combinations of the two reference and two quadrate quantities. With a typical electrical flow graph less than 20 per cent of the possible transmissions are non-zero and therefore the check for zero terms reduces the time taken by any program by a factor of five, or more.

The third procedure in Appendix (b) is concerned with setting up the numerical values, or expressions, for the main body of the amplifier flow graph. The nodes of the flow graph are numbered as in Fig. 7.

Those branches given as solid lines in Fig. 7 are set up in procedure 80, since these have the same location and node numbers for all required operations on the flow graph. The branches given as dashed lines in Fig. 7 may, or may not, be used, depending on the relationship required. For the voltage gain through the amplifier node 1 is on the left-hand side and node 10 at the right-hand side. But to find the input admittance (the relationship between V_i and i_{11}) node 10 may be moved to its upper left position. Or, to find the output admittance (which is the relationship between V_o and i_{22}) node 1 is moved to the lower right-hand side and node 10 joined directly to node 9. It is at this stage that the flow graph becomes a valuable aid in writing a correct and logical computer program. The numerical transmissions associated with these movable branch locations are put into the program where appropriate.

Since the voltage gain has to be computed a number of times procedure 90 was written to perform this operation. For this purpose it is assumed that a family of ten similar transistors is available, with manufacturer's type numbers from 1210 to 1219, where the gain/current characteristic of the lowest-priced transistor, in this family, is as in Table 1. The current gain increases by 10 per cent for each successively higher numbered transistor in the range. So NT can have any value from 0 to 9 within the program, and the common-emitter current gain of any transistor, at the N^{th} step of emitter current, is given by:

$$G = H(N) \times (1{\cdot}1)^{NT}$$

Two versions of a main program were written and the first is given in Appendix (c). A numerical gain specification, known as the Gain Aim (GA) is read into the computer, after the set of current and gain values (Table 1). The main program then searches for the least value of emitter current and the lowest numbered transistor type which will satisfy this gain specification. If this is an impossible specification, within the given range of devices and current, then an appropriate message is printed. When the gain aim can be met then full details of transistor type number, emitter current, voltage gain and phase angle, input admittance and impedance, and output admittance and impedance are printed. It should be noted that

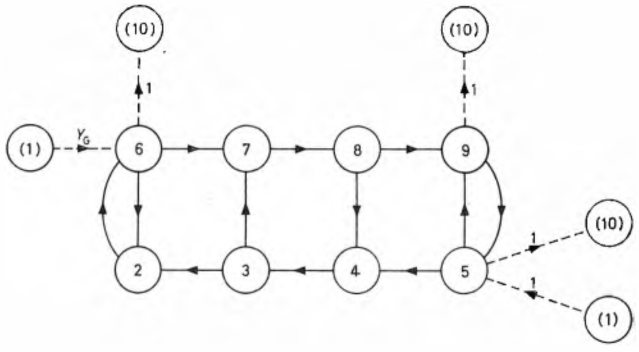


Fig. 7. Node numbering for the amplifier flow graph

in this program minimum emitter current was taken as the primary consideration. This would be the case if a maximum packing density of amplifier circuits within a small unit was a design requirement. Some typical results from this design program are printed in Appendix (e).

Cost Minimization

A second main program was written to approach the design problem from a different point of view, but using the same program procedures, and this is given in Appendix (d). This program also requires a numerical gain aim, which is read in after the quantities from Table 1. In this program the transistor with the highest gain, which is type number 1210 (corresponding to $NT = 9$) is chosen first and the voltage gain is computed for the minimum current value of $C(1) = 0{\cdot}1$mA. If this does not give the required gain then current steps from 2 to 10 are used in turn. It is necessary to check through the current/gain steps one by one, since it is by no means certain that the largest emitter current will give the highest gain. The I.B.M. 1620 has a basic store cycle time of 20μsec and consequently takes about 22sec to execute procedure 90 (which includes procedures 60, 70 and 80) once. In other words, the computer requires less than two minutes to check whether it is impossible to attain a certain gain aim. With a medium-sized computer these operations could be completed in less than one-tenth of the time.

If a gain aim is obtainable then the program computes an estimated price for the design, based on the formula:

$$P = 10 + 8T + 2{\cdot}5\,C(N) + 0{\cdot}2\,C(N)^2$$

In this cost function the figure of 10 represents the fixed cost of the components, other than the transistors. T represents the cost of the two transistors, where it is reckoned that the transistor prices go up within the family by steps of 20 per cent, so that $T = (1{\cdot}2)^{NT}$. The price formula also includes an estimate, of $2{\cdot}5\,C(N)$, for the cost of standing current, which affects the size of the power supply, or the battery drain, and a final term of $0{\cdot}2\,C(N)^2$ to put a price on power dissipation.

Having completed this computation for one transistor the program proceeds to the transistor with the next type number downwards and repeats this process until no other transistor, at any of the ten possible current settings, can meet the gain aim. By checking over the printed results, of which one set is in Appendix (e) it is possible to see at a glance which is the most suitable amplifier design for a particular requirement.

A useful by-product of this program is that it is possible quickly to produce a set of graphs relating transistor type and emitter current for a given gain aim. So far as is known graphs of this type have not been previously published (probably because of the computation effort

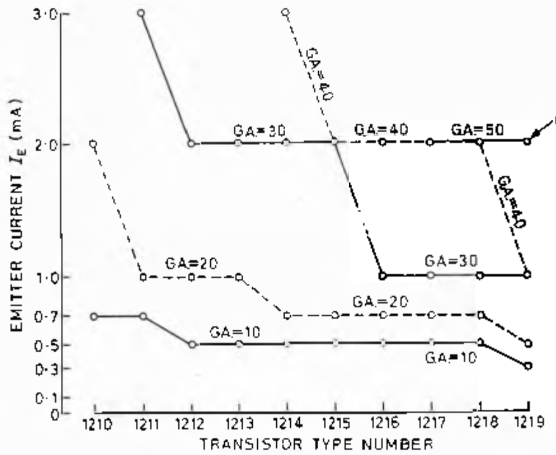


Fig. 8. Graphs relating emitter current, transistor type and gain aim

involved). One such set of graphs, for the amplifier circuit considered in this article, is drawn as Fig. 8.

A more complicated cost function could be included in the program, together with certain practical limitations on the amplifier design, and the details of an optimum solution arrived at by the methods of linear programming. However, this extension of the program would require a digital computer with a larger store than was available, if a flow graph of worthwhile size (say, ten or more nodes) was to be considered.

Conclusions

The example of an amplifier design shows how the methods of flow graphs may be combined with an automatic programming language to produce worthwhile results. The procedures given in this article are by no means limited to electronic circuits, since meaningful flow graphs can be drawn for many physical systems, in such fields as mechanical engineering, operational research and economics[6]. The construction of a flow graph for any one of these systems avoids the necessity of writing down the detailed system equations and is the significant step between the problem and its 'electronic model' in the computer.

Acknowledgments

The author wishes to thank Dr. J. G. Byrne (Trinity College, Dublin) and Mr. G. P. Coverley (Northern Electric Company, Ottawa) for their helpful advice and encouragement. The permission of the Pergamon Press to reproduce Fig. 4 is also acknowledged.

Appendix

The programs in this appendix were written in Fortran (U.T.O. version) for the I.B.M. 1620 computer. Only minor changes are necessary for other versions of Fortran.

(a) Simple Program for Real Transmissions.

(b) Procedures for Complex Flow Graphs.

(c) First Main Program.

(d). Second Main Program.

(e) Typical results from two main programs.

(a) Simple Program for Real Transmissions

```
C REDUCTION PROGRAM FOR REAL FLOW GRAPH.
DIMENSION T(30,30)
PRINT50
50 FORMAT(34H TRANSMISSION THROUGH A FLOW
GRAPH)
NG=1
1 PRINT51, NG
51 FORMAT(/15H FLOW GRAPH NO.,I3)
READ40,NN
40 FORMAT(I2)
NA=NN-1
DO 2 I=1,NA
DO 2 J=2,NN
2 READ41,T(I,J)
41 FORMAT(F17·8)
DO 3 L=2,NA
LA=L+1
D=1./(1.-T(L,L))
DO 3 M=LA,NN
T(1,M)=T(1,L)*T(L,M)*D+T(1,M)
IF(LA-NN)4,5,5
4 DO 3 K=LA,NA
3 T(K,M)=T(K,L)*T(L,M)*D+T(K,M)
5 PRINT52,T(1,NN)
52 FORMAT(5X,22H THROUGH TRANSMISSION=,F14·6)
NG=NG+1
GO TO 1
END
TRANSMISSION THROUGH A FLOW GRAPH
FLOW GRAPH NO. +1
THROUGH TRANSMISSION=      +7·500000
FLOW GRAPH NO. +2
THROUGH TRANSMISSION= +8000·000000
```

(b) Procedures for Complex Flow Graphs

```
C COMPLEX FLOW GRAPH REDUCTION.
BEGIN PROCEDURE 60
NA=NN-1
DO 61 L=2,NA
LA=L+1
RL=1.-R(L,L)
DD=RL*RL+Q(L,L)*Q(L,L)
RD=RL/DD
QD=Q(L,L)/DD
DO 61 M=LA,NN
K=1
EXECUTE PROCEDURE 70
IF(LA-NA)62,62,63
62 DO 61 K=LA,NA
EXECUTE PROCEDURE 70
61 CONTINUE
63 AN=SQRT(R(1,NN)*R(1,NN)+Q(1,NN)*Q(1,NN))
BN=1./AN
PH1=(180./.14)*ATAN(Q(1,NN)/R(1,NN))
PH2=0·0-PH1
END PROCEDURE 60
C COMPLEX OPERATIONS.
BEGIN PROCEDURE 70
IF(R(K,L)-0.0)71,72,71
71 IF(R(L,M)-0.0)73,72,73
73 RR=R(K,L)*R(L,M)-Q(K,L)*Q(L,M)
QQ=R(K,L)*Q(L,M)+Q(K,L)*R(L,M)
R(K,M)=R(K,M)+RD*RR-QD*QQ
Q(K,M)=Q(K,M)+RD*QQ+QD*RR
72 END PROCEDURE 70
C NUMERICAL BRANCH TRANSMISSIONS.
BEGIN PROCEDURE 80
NA=NN-1
DO 81 I=1,NA
DO 81 J=2,NN
R(I,J)=0·0
81 Q(I,J)=0·0
R(2,6)=-0·0025
Q(2,6)=-0·0005
R(3,2)=0·007
Q(3,2)=0·003
R(3,7)=0·004
```

```
Q(3,7)=0·001
R(4,3)=1·0
R(5,4)=0·005
Q(5,4)=0·02
R(5,9)=0·0001
Q(5,9)=0·0015
R(6,2)=500·/C(N)
Q(6,2)=120·/C(N)
R(6,7)=−0·94*G
Q(6,7)=−0·35*G
R(7,8)=−1·0
R(8,4)=25·/C(N)
Q(8,4)=6·/C(N)
R(8,9)=−G/(1·+G)
Q(8,9)=−0·02*G/(1·+G)
R(9,5)=−500.
END PROCEDURE 80
C VOLTAGE GAIN.
BEGIN PROCEDURE 90
G=H(N)*(1·1**NT)
EXECUTE PROCEDURE 80
R(1,6)=0·0025
Q(1,6)=0·0005
R(5,10)=1·0
EXECUTE PROCEDURE 60
END PROCEDURE 90
```

(c) First Main Program

```
C MAIN PROGRAM 1.
DIMENSION R(10,10),Q(10,10),C(10),H(10)
PRINT50
50 FORMAT(27H TWO-STAGE AMPLIFIER DESIGN)
NN=10
DO 1 N=1,10
1 READ40,C(N),H(N)
40 FORMAT(F5.1)
2 READ40,GA
DO 10 N=1,10
NT=9
EXECUTE PROCEDURE 90
IF(AN−GA)5,3,4
4 NT=5
EXECUTE PROCEDURE 90
IF(AN−GA)8,3,6
6 NT=0
7 EXECUTE PROCEDURE 90
IF(AN−GA)8,3,3
8 NT=NT+1
GO TO 7
5 IF(N−10)10,9,9
10 CONTINUE
9 PRINT51,GA
51 FORMAT(//8H GAIN OF,F6·1,26H NOT
    OBTAINABLE, REDUCE GA)
GO TO 2
3 NT=NT+1210
PRINT52,NT,GA
52 FORMAT(//6X,20H TRANSISTOR TYPE NO.,I5,5X,
    10H GAIN AIM=,F6·1)
PRINT53,C(N)
53 FORMAT(/17H EMITTER CURRENT=,F5·1,3H MA)
PRINT54,AN,PH1
54 FORMAT(/14H VOLTAGE GAIN=,F5·1,3H AT
    ,F5·1,5H DEG.)
EXECUTE PROCEDURE 80
R(1,6)=0·0025
Q(1,6)=0·0005
R(6,10)=1·0
EXECUTE PROCEDURE 60
PRINT55,R(1,10),Q(1,10)
55 FORMAT(/18H INPUT ADMITTANCE=,F8·6,F8·6,
    5HJ MHO)
PRINT56,BN,PH2
56 FORMAT(17H INPUT IMPEDANCE=,F7·0,7H OHM
    AT,F5·1,5H DEG.)
EXECUTE PROCEDURE 80
R(1,5)=1·0
R(9,10)=1·0
EXECUTE PROCEDURE 60
PRINT57,R(1,10),Q(1,10)
57 FORMAT(/19H OUTPUT ADMITTANCE=,F8·6,
    F8·6,5HJ MHO)
PRINT58,BN,PH2
58 FORMAT(18H OUTPUT IMPEDANCE=,F7·0,7H
    OHM AT,F5·1,5H DEG.)
GO TO 2
```

(d) Second Main Program

```
C MAIN PROGRAM 2.
DIMENSION R(10,10),Q(10,10),C(10),H(10)
PRINT50
50 FORMAT(27H TWO-STAGE AMPLIFIER PRICES)
NN=10
DO 1 N=1,10
1 READ40,C(N),H(N)
40 FORMAT(F6·1)
8 READ40,GA
PRINT51,GA
51 FORMAT(//10X,10H GAIN AIM=,F6·1)
NT=9
N=1
4 EXECUTE PROCEDURE 90
IF(AN−GA)2,3,3
2 N=N+1
GALLEY 2
IF(N−10)4,4,5
5 IF(NT−9)6,7,7
6 PRINT52,GA
52 FORMAT(/37H NO OTHER AMPLIFIER WILL GIVE GAIN
    OF, F6·1)
GO TO 8
7 PRINT53,GA
53 FORMAT(/8H GAIN OF, F6·1,26H NOT OBTAINABLE,
    REDUCE GA)
GO TO 8
3 P=10·+8·*(1·2**NT)+2·5*C(N)+0·2*C(N)*C(N)
NC=NT+1210
PRINT54,NC,C(N)
54 FORMAT(/20H TRANSISTOR TYPE NO., I5,17H EMITTER
    CURRENT=,F5·1,3H MA)
PRINT55,AN,PH1
55 FORMAT(5X,14H VOLTAGE GAIN=,F5·1,3H AT,F5·1,
    5H DEG.)
PRINT56,P
56 FORMAT(20H PRICE OF AMPLIFIER=,F6·1,10H SHILLINGS)
NT=NT−1
IF(NT−0)6,4,4
```

(e) Typical Results from Two Main Programs

```
TWO-STAGE AMPLIFIER DESIGN

GAIN OF+200·0 NOT OBTAINABLE, REDUCE GA
        TRANSISTOR TYPE NO.+1217     GAIN AIM= +50·0
EMITTER CURRENT= +2·0 MA
VOLTAGE GAIN=+52·5 AT−14·9 DEG.
INPUT ADMITTANCE=+·001562+·000011J MHO
INPUT IMPEDANCE=  +640· OHM AT  −·4 DEG.
OUTPUT ADMITTANCE=+·000747+·000910J MHO
OUTPUT IMPEDANCE=  +849· OHM AT−50·6 DEG.
        TRANSISTOR TYPE NO.+1216     GAIN AIM=+30·0
EMITTER CURRENT= +1·0 MA
VOLTAGE GAIN=+31·2 AT−20·3 DEG.
```

```
INPUT ADMITTANCE=+·001114–·000073J MHO
INPUT IMPEDANCE= +895· OHM AT +3·7 DEG.
OUTPUT ADMITTANCE=+·000744+·000906J MHO
OUTPUT IMPEDANCE= +852· OHM AT–50·6 DEG.
TWO-STAGE AMPLIFIER PRICES
        GAIN AIM=+200·0
GAIN OF+200·0 NOT OBTAINABLE, REDUCE GA
        GAIN AIM= +60·0
TRANSISTOR TYPE NO.+1219 EMITTER CURRENT= +2·0 MA
    VOLTAGE GAIN=+63·5 AT–15·1 DEG.
PRICE OF AMPLIFIER= +57·0 SHILLINGS
NO OTHER AMPLIFIER WILL GIVE GAIN OF +60·0
        GAIN AIM= +50·0
TRANSISTOR TYPE NO.+1219 EMITTER CURRENT= +2·0 MA
    VOLTAGE GAIN=+63·5 AT–15·1 DEG.
PRICE OF AMPLIFIER= +57·0 SHILLINGS
TRANSISTOR TYPE NO.+1218 EMITTER CURRENT= +2·0 MA
    VOLTAGE GAIN=+57·8 AT–15·0 DEG.
PRICE OF AMPLIFIER= +50·1 SHILLINGS
TRANSISTOR TYPE NO.+1217 EMITTER CURRENT= +2·0 MA
    VOLTAGE GAIN=+52·5 AT–14·9 DEG.
PRICE OF AMPLIFIER= +44·4 SHILLINGS
NO OTHER AMPLIFIER WILL GIVE GAIN OF +50·0
```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An Iterative Method of Solving Multiple Mesh Problems

By N. M. Morris*, B.Sc., A.M.I.E.E., A.M.I.E.R.E.

An iterative method of solving multiple mesh problems is described. This is based on Taylor's series and is applicable to any linear a.c. or d.c. network and to some non-linear ones. It offers a useful alternative approach to the solution of multiple mesh networks where a simulator is not available.

(Voir page 776 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

WHERE multiple-mesh problems have to be solved the difficulties in manipulating the equations increase rapidly with the number of meshes. This is further aggravated by the drooping or rising voltage-current characteristics of the sources of supply within the network.

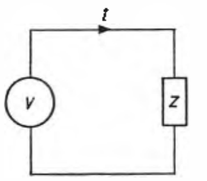

Fig. 1. Single-loop circuit

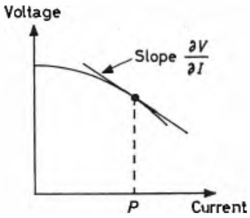

Fig. 2. Generator-potential-current characteristic

Alternative methods of solution include the use of iterative procedures in which an approximate solution, based on experience, is first used. The resulting error allows the first approximation to be modified, giving a more accurate solution; after a few iterations the error usually falls to a very low value. In this article an iterative method based on Taylor's series is developed which is applicable to direct and alternating current networks.

A Single-Loop Problem

An elementary loop is shown in Fig. 1 in which the voltage V and impedance Z are known, but the current I is unknown. The shape of the voltage-current characteristic is also known. Using this circuit it is proposed to develop a technique that can be applied to multi-mesh problems.

Since the current is not known, an assumed value, I, is used in the calculations such that:

$$ZI - V = f(Z, I) \neq 0 \quad \ldots\ldots\ldots\ldots (1)$$

If the selected value of I is not correct, Kirchhoff's second law is not satisfied and $f(Z, I)$ cannot be zero. If a 'correcting' current, x, is now allowed to flow in the loop, the new current I' is given by:

$$I' = I + x$$

It is assumed that I' makes equation (1) zero. For convenience this is rewritten:

$$f(Z, I') = f(Z, I + x) = Z(I + x) - V \quad \ldots\ldots (2)$$

Expanding $f(Z, I')$ as a Taylors series:

$$f(Z, I+x) = f(Z, I) + x\,(\partial f/\partial I)\,(Z, I+x) + (x^2/2!)\,(\partial^2 f/\partial I^2)\,(Z, I + x) + \ldots..$$

If the initial value of I is nearly correct, terms greater than x may be neglected, giving:

$$f(Z, I+x) = f(Z, I) + x\,\partial(ZI+Zx-V)/\partial I$$
$$= f(Z, I) + xZ - x\partial V/\partial I \quad \ldots\ldots\ldots\ldots (3)$$

Since it is assumed that current $(I + x)$ makes equation (2) zero, it must also make equation (3) zero,

$$\therefore\ f(Z, I) + x(Z - \partial V/\partial I) = 0$$

and correcting current

$$x = -\frac{f(Z, I)}{Z - \partial V/\partial I}$$

In practice the slope of the voltage-current characteristic is not zero at the assumed operating point P (Fig. 2) and must be taken into account.

The value of the correcting current is inserted into equation (2) to see if it makes $f(Z, I + x)$ zero. If this is not the case a new value of correcting current x' has to be evaluated giving a new loop current I'' where:

$$I'' = I' + x'$$

and:

$$x' = -\frac{f(Z, I + x)}{Z - \partial V/\partial I'}$$

$\partial V/\partial I'$ is the slope of the generator characteristic at the new operating point. The process is repeated until the calculated correcting current is very small or zero.

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Examples of Application

Direct Current Circuit

The source (Fig. 3) has an open-circuit e.m.f. of 10V and a uniformly drooping characteristic over the load range of $\partial V/\partial I = -1\text{V/A}$.

A circuit current must be assumed and the unlikely value of 2A is chosen, giving a generator terminal voltage of 8V. Inserting these values into equation (1)

$$f(Z, I) = IR - V = (2 \times 9) - 8 = 10\text{V}$$

$$x = -\frac{f(Z, I)}{Z - \partial V/\partial I} = -\frac{10}{9+1} = -1\text{A}$$

$$I' = I + x = 2 - 1 = 1A$$

With this value of current the generator terminal voltage is 9V.

$$f(Z, I') = R(I + x) - V = 9 - 9 = 0$$

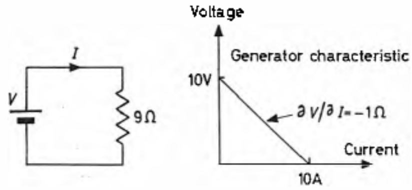

Fig. 3. Elementary direct current circuit

The value of x' is thus zero and the circuit current is therefore 1A.

Alternating Current Circuit

In Fig. 4 let $V = \sqrt{2}\text{V}$ and $\partial V/\partial I$ be zero. Assume initially that a current $(1 + \text{j}0)$ amperes circulates.

$$f(Z, I) = (1 + \text{j}1) - 1{\cdot}414 = -0{\cdot}414 + \text{j}1$$

$$x = -\frac{f(Z, I)}{Z - \partial V/\partial I} = \frac{0{\cdot}414 - \text{j}1}{1 + \text{j}1} = -0{\cdot}293 - \text{j}\,0{\cdot}707$$

$$I' = I + x = 0{\cdot}707 - \text{j}\,0{\cdot}707 = 1\angle{-45^\circ}$$

$$f(Z, I') = 1\angle{-45^\circ} \,.\, \sqrt{2}\angle{45^\circ} - \sqrt{2} = 0$$

The current in the circuit is $1\angle{-45^\circ}$ A.

Multi-Mesh Networks

A section of a multi-mesh network is shown in Fig. 5 in which assumed currents $I_1, I_2, \dots I_n$, etc., circulate round meshes $1, 2, \dots n$, etc. The self-impedance of mesh n is Z_{nn} and the mutual impedance between meshes m and n is Z_{mn}.

The mesh equations are:

Mesh 1

$$-I_1Z_{11} + I_2Z_{12} - \dots - I_nZ_{1n} - \dots - V_1 = a_1$$

Mesh 2

$$I_1Z_{21} + I_2Z_{22} - \dots - I_nZ_{2n} - \dots - V_2 = a_2$$

Mesh n

$$-I_1Z_{n1} - I_2Z_{n2} - \dots + I_nZ_{nn} - \dots - V_n = a_n \qquad \dots\dots (4)$$

Correcting currents $x_1, x_2, \dots x_n$, etc., are allowed to flow in meshes $1, 2, \dots n$, etc., resulting in new mesh currents $I_1' = I_1 + x_1$, $I_2' = I_2 + x_2, \dots I_n' = I_n + x_n$, etc. Substituting these into equation (4) and expanding as a Taylors series, neglecting second order and higher terms gives:

$$\begin{aligned} f(I_1', I_2', \dots I_n', \dots) &= f(I_1 + x_1, I_2 + x_2, \dots I_n + x_n, \dots) \\ &= a_1 + x_1(\partial a_1/\partial I_1) + x_2(\partial a_1/\partial I_2) + \dots + x_n(\partial a_1/\partial I_n) + \dots \\ &= a_1 + x_1(Z_{11} - \partial V_1/\partial I_1) - x_2Z_{12} - \dots - x_nZ_{1n} - \dots \qquad \dots\dots (5) \end{aligned}$$

For other meshes the resulting equations are:

Mesh 2

$$a_2 - x_1Z_{21} + x_2(Z_{22} - \partial V_2/\partial I_2) - \dots - x_nZ_{2n} - \dots \qquad \dots\dots (6)$$

Mesh n

$$a_n - x_1Z_{n1} - x_2Z_{n2} - \dots + x_n(Z_{nn} - \partial V_n/\partial I_n) - \dots \qquad \dots\dots (7)$$

Assuming that reasonable estimates of current magnitudes have been made, and that

$$I_1Z_{11} \gg I_2Z_{12}, \dots I_nZ_{1n}, \text{ etc.}$$
$$I_2Z_{22} \gg I_1Z_{21}, \dots I_nZ_{2n}, \text{ etc.}$$
$$I_nZ_{nn} \gg I_1Z_{n1}, \dots, \text{ etc.}$$

and the correcting currents make equations (5), (6) and (7) zero, they simplify to respectively

$$a_1 + x_1(Z_{11} - \partial V_1/\partial I_1) = 0$$
$$a_2 + x_2(Z_{22} - \partial V_2/\partial I_2) = 0$$
$$a_n + x_n(Z_{nn} - \partial V_n/\partial I_n) = 0$$

The correcting flow for the n^{th} mesh of a network is

$$x_n = -\frac{a_n}{Z_{nn} - \partial V_n/\partial I_n}$$

In many cases the conditions outlined are not satisfied, but this does not invalidate the general procedure. The net result in such a case is that more iterations must be made before the final solution is obtained.

The resulting currents $I_1', I_2', \dots I_n', \dots$ etc., can be calculated and $f(I_1', I_2', \dots I_n', \dots)$ evaluated. If this is not zero, new correcting flows $x_1', x_2', \dots x_n', \dots$ etc., can be determined, where

$$x_n' = -\frac{a_n'}{Z_{nn} - \partial V_n/\partial I_n'}$$

giving a new current in the n^{th} mesh of

$$I_n'' = I_n' + x_n'$$

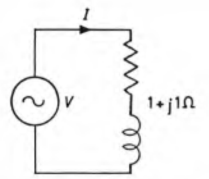

Fig. 4. Elementary alternating current circuit

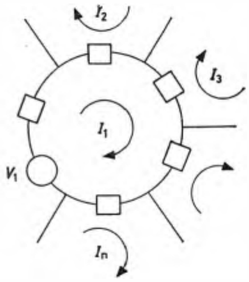

Fig. 5. General multiple-mesh problem

The process is repeated until the correcting current is very small or zero.

Non-Linear Networks

The above procedure can be extended to non-linear networks with a voltage-current relationship of the form

$$V = RI^{\alpha}$$

Inserting this relationship into the network equations and expanding as a Taylors series, adopting the assumptions already made, the correcting flow for the p^{th} mesh in a q mesh network is:

$$x_p = -a_p/K_{pp}$$

where

$$K_{pp} = \sum_{n=1}^{n=q} \alpha Z_{pn} \,|\, I_{pn} \,|^{\,\alpha-1} - \partial V_p/\partial I_p$$

Conclusion

The method outlined is applicable to any linear electrical network and to some non-linear ones. It offers an alternative approach to the solution of multiple-mesh networks where a simulator is not available. One advantage of iterative methods of solution is that a mistake at some point in the calculation will not alter the accuracy of the final solution.

A Transistorized Ratio Amplifier Employing the Principle of Elapsed Time Computation

By A. A. Cremin*, M.Eng.Sc. and M. C. Sexton*, M.Sc., Ph.D., A.M.I.E.E., A.Inst.P.

A description is given of an analogue circuit suitable for obtaining the ratio of two time-varying input signals. The method involves the use of elapsed time (time-base) circuits to generate sampling frequencies considerably higher than the input signal frequencies thereby producing an essentially continuous output signal proportional to the input ratio.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

In connexion with experimental studies of plasma resonances in d.c. discharges using a frequency-swept backward wave oscillator it was necessary to design a ratio amplifier of the elapsed-time computation type[1,2] with a sampling time of less than 50μsec and an overall accuracy of the order of 1 per cent. The circuit to be described meets these requirements satisfactorily using only eight transistors.

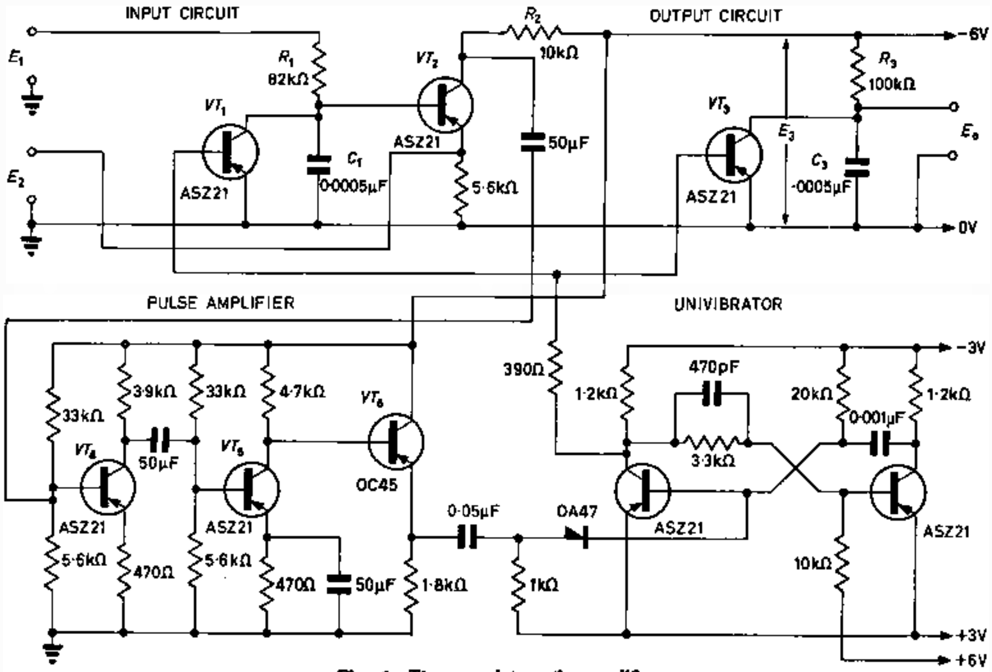

Fig. 1. The complete ratio amplifier

Theory of Operation

The complete circuit diagram is shown in Fig. 1 with the following cycle of operation: assuming that the univibrator has just been triggered, a negative pulse is applied to the bases of transistors VT_1 and VT_3 ensuring that the voltages across capacitors C_1 and C_3 fall approximately to zero. At the end of this negative pulse the bases are driven positive, cutting off the two transistors, and C_1 and C_3 begin to charge.

Let this instant be called $t = 0$. If E_1 and E_2 are the voltages whose ratio is required (assumed constant over one cycle of operation) and E_3 is a fixed voltage across the network R_3C_3, then:

$$V_1 = E_1\,[1 - \exp(-t/R_1C_1)] \quad \ldots\ldots\ldots\ (1)$$

and

$$V_3 = E_3\,[1 - \exp(-t/R_3C_3)] \quad \ldots\ldots\ldots\ (2)$$

where V_1 and V_3 are the instantaneous voltages across C_1 and C_3 respectively.

Assuming that E_1 and E_2 are both negative ($|E_1| > |E_2|$) transistor VT_2 is cut off until V_1 becomes equal to E_2. VT_2 then begins to conduct and the resultant pulse developed across R_2 is amplified in pulse amplifier $VT_{4,5,6}$ and used to trigger the univibrator which returns the circuit to its initial condition.

Let the peak value of V_3 be E_0. The peak value of V_1 is, of course, E_2. Inserting these values in equations (1) and (2) and eliminating t:

$$1 - (1 - (E_0/E_3))^{\alpha} = (E_2/E_1) \quad \ldots\ldots\ldots\ (3)$$

where $\alpha = R_3C_3/R_1C_1$.

In the particular case where $R_1C_1 = R_3C_3$, equation (3) reduces to:

$$E_0 = (E_2/E_1)\,E_3 \quad \ldots\ldots\ldots\ldots\ldots\ (4)$$

Thus, the amplitude E_0 of the sawtooth output voltage is directly proportional to the ratio of the input voltages E_1 and E_2 provided that E_3 is fixed and E_1 does not change appreciably during each sampling period.

Sources of Error

The error due to unequal time-constants increases with the ratio E_2/E_1. If, for example, this ratio exceeds 0·4, the error becomes greater than 1 per cent when the time-constants differ by 5 per cent. Thus, the upper limit of

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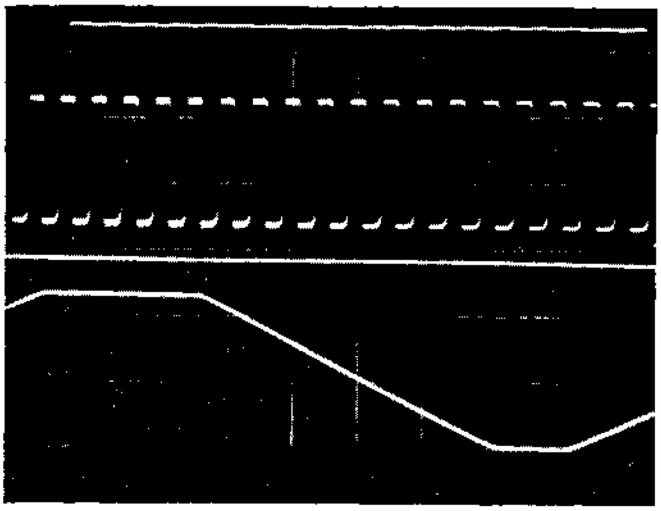

Fig. 2. Test waveforms E_1 and E_2

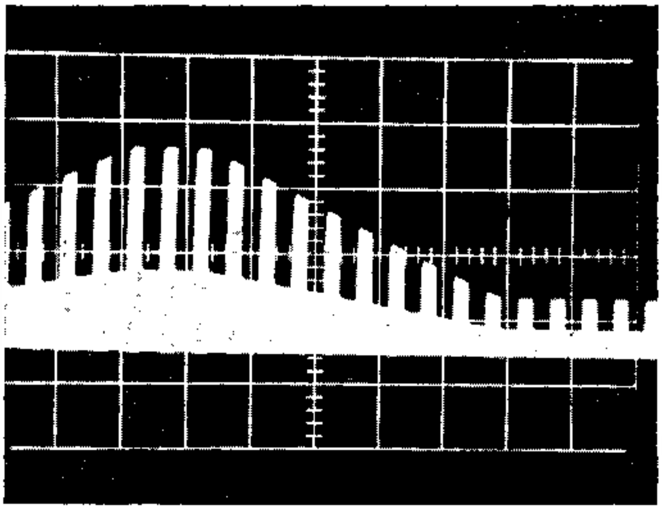

Fig. 3. The ratio E_2/E_1 obtained from the time-varying signals of Fig. 2

the ratio E_2/E_1 which can be handled is determined by the accuracy with which the two time-constants can be matched. It should be noted that the output impedance of the source of E_1 must be included in R_1.

As the ratio E_2/E_1 becomes smaller the error discussed above becomes insignificant. At the same time, however, the associated reduction of the sampling time introduces another difficulty since for a given circuit a fixed delay-time (Δt) exists between the instant at which VT_2 conducts and that at which VT_1 and VT_3 conduct. If t is the sampling time the error from this source will be $\Delta T/t \times 100$ per cent for small values of E_2/E_1. In the original circuit, using OC45 transistors throughout, the delay time was found to be almost 2μsec. Hence, for a 1 per cent error the minimum allowable sampling time approached 200μsec. Since sampling time increases with ratio the range of sampling times extended up to 2msec in practice.

The subsequent use of ASZ21 transistors reduced this time-delay to approximately 0·5μsec. This gave a range of sampling time extending from 50 to 500μsec. Since these sampling times were still a factor of ten longer than desired, the error introduced by the time-delay was examined in greater detail.

When the time delay Δt is taken into account equation (4) becomes:

$$E_0 = E_2/E_1 \,.\, E_3\,[1 - (\Delta t/R_3C_3)] + (\Delta t/R_3C_3)\,.\,E_3 \qquad \ldots\ldots\ldots\ldots (5)$$

assuming $\Delta t \ll R_3C_3$. Thus the linearity of the relationship between E_0 and the ratio E_2E_1 is not affected by the presence of Δt. The constant zero-error $\Delta t/R_3C_3\,.\,E_3$ is relatively insignificant in practice since it can be eliminated on a c.r.o. presentation by merely drawing a new zero line. Electrical subtraction may be used if the output is driving a further stage. With the 5 to 50μsec range of sampling times used by the authors the zero-error was about 10 per cent of the smallest output voltage encountered.

In order to cater for 5μsec sampling times the univibrator was designed with a recovery time of 2·5μsec.

In the derivation of equation (4) it was assumed that a pulse large enough to trigger the univibrator is obtained when the base and emitter voltages of transistor VT_2 become equal. In practice, of course, a small but finite voltage difference is required. For a given sampling time this imposes a lower limit on the value of E_2 than can be accurately handled. In the circuit of Fig. 1 the error from this sourse is negligible for $E_2 > 0{\cdot}2$V.

Finally, the effect of leakage currents in transistors VT_1, VT_2 and VT_3 was minimized by selecting low-leakage transistors for these positions.

Performance

The input voltages shown in Fig. 2 were used to assess the performance of the circuit. Here E_1 is the square wave with a repetition frequency of 500c/s and E_2 the ramp function repetitive at 20c/s, with peak amplitudes of 3 and 0·6V respectively. The range of ratios covered by these waveforms extends from 0·05 to 0·5, with corresponding sampling times varying from approximately 5 to 50μsec. With a univibrator pulse length of 10μsec, the sampling frequencies corresponding to the above times are 65 and 17kc/s respectively.

The ratio of the above time functions is shown in Fig. 3. It is clear that the linearity of the relationship between E_0 and the ratio E_2/E_1 is better than 1 per cent.

Conclusions

The accuracy of the amplifier holds as long as E_2/E_1 lies in the range $0{\cdot}05 < E_2/E_1 < 0{\cdot}5$, with the instantaneous value of E_2 within the limits $0{\cdot}2 < E_2 < 2$V negative. Under these conditions the maximum sampling time is 50μsec but the lower limit of 0·05 may, of course, be extended at the expense of sampling time. To maintain the input voltage ratios between the above limits, the use of a pre-amplifier with variable gain will clearly extend the operating range without loss of accuracy. In addition, envelope detection of the output would enable the circuit to be used as a standard analogue computing element.

Acknowledgments

The authors wish to thank Professor C. T. G. Dillon for making available the facilities of the Electrical Engineering Department. The financial assistance of the English Electric Valve Co. Ltd, Chelmsford, through a post-graduate award to one of the authors (A.A.C.) together with technical discussions with Mr. G. O. Chalk of the Company are gratefully acknowledged. The work was also supported, in part, by the Air Force Office of Scientific Research under Grant EOAR 65-46 through the European Office of Aerospace Research, United States Air Force.

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Negative Output Resistance in a Regulated Power Supply

By J. Diamond*

A negative output resistance is sometimes desirable in a regulated power supply. It might be used to cancel the resistance of leads, or some component in series with the load, where the usual solution—remote sampling of the load voltage—is inapplicable.
The design of such a supply is discussed. It is shown that a constant negative output resistance may be obtained, independent of output voltage. The temperature coefficient of the output resistance may also be controlled.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

In certain cases it is desirable to achieve a negative output resistance in a regulated power supply. This negative resistance might be used to cancel the resistance of leads or of some component in series with the load. While the usual solution for this problem is remote sampling of the load voltage, in some cases the point which would have to be sampled is inaccessible, or cannot take the loading of the sampling network. It must be recognized, however, that cancellation by means of negative resistance is a form of open-loop compensation, not of feedback. Thus it cannot deal with changes in the external resistance which it should compensate, unless these are predictable (such as changes due to ambient temperature, which occur in accordance with a known temperature coefficient).

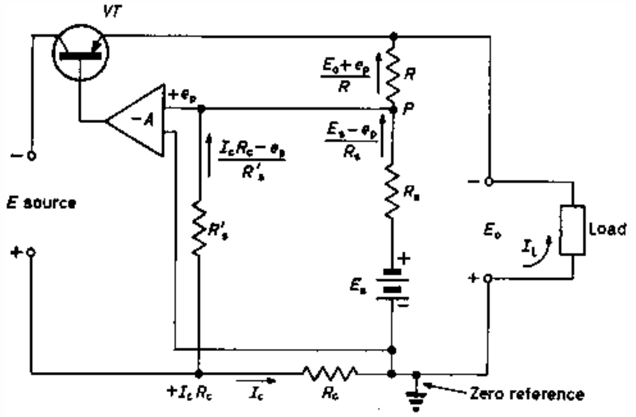

Fig. 1. Regulated power supply with negative output resistance

Circuit

Fig. 1 shows the control circuit for a regulated supply with provision for negative output resistance. With R_s' omitted and R_c short-circuited, the circuit is a conventional regulator with a separate reference voltage E_s. R_c and R_s' are introduced to inject a component proportional to load current into the input of the control amplifier A. The polarity of this component is such as to increase the output as the load current increases—the circuit permits the output, the reference supply, and the control amplifier to have a common earth. The series element is shown as a transistor (VT), but the scheme applies to any type of regulator.

Analysis

It is assumed that the control amplifier A provides sufficient loop gain to keep the signal at its own input (point P) zero. To evaluate the effect of d.c. drift in A, however, the voltage at P is shown as e_p rather than zero. A is also assumed to take no input signal current. Although the input impedance of A may be rather low in the case of a transistor amplifier, this assumption is justified, because it is assumed that e_p is zero except for drift effects. Summing the currents at point P:

$$\frac{I_cR_c - e_p}{R_s'} + \frac{E_s - e_p}{R_s} = \frac{E_o + e_p}{R} \quad \ldots\ldots (1)$$

where:

$$I_c = I_L + \frac{E_s - e_p}{R_s} \quad \ldots\ldots\ldots\ldots (2)$$

solving for E_o:

$$E_o = E_s\,(R/R_s)\,(1 + (R_c/R_s')) + \frac{R_cR}{R_s'}\,I_L - e_p\left[1 + (R/R_s) + (R/R_s') + \frac{R_cR}{R_sR_s'}\right] \quad \ldots\ldots (3)$$

The presence of an I_L term with a positive coefficient shows that the output voltage will increase as I_L increases, indicating a negative output resistance whose magnitude is this coefficient. Using the symbol R_o for this (positive) magnitude:

$$R_o = R_cR/R_s' \quad \ldots\ldots\ldots\ldots (4)$$

Substituting equation (4) into equation (3):

$$E_o = (E_s/R_s)\,(R + R_o) + R_oI_L - e_p\left(1 + \frac{R + R_o}{R_s} + R_o/R_c\right) \quad \ldots\ldots\ldots\ldots (5)$$

In equation (5) the output voltage is separated into three components: an open-circuit voltage, a rise due to R_o, and a drift term.

The open-circuit voltage is:

$$E_{oc} = (E_s/R_s)\,(R + R_o) \quad \ldots\ldots\ldots\ldots (6)$$

The drift term, using equation (6), is:

$$E_{drift} = -e_p\,(1 + (R_o/R_c) + (E_{oc}/E_s)) \quad \ldots\ldots (7)$$

Adjustment of Output Voltage

As seen from equation (6), the output voltage may be adjusted by varying E_s, R_s or R. Varying R is undesirable because it will directly effect R_o (equation (4)), leaving a choice of E_s or R_s. R_s may be used when the non-linear variation of E_o with R_s is acceptable, for example in a supply with a limited output range. It is also possible to use a non-linear rheostat, or to use decade control where the decades are arranged to give equal steps of conductance. If linear control with a linear potentiometer is

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desired, however, the only possibility is to vary E_s. This case is important because of the convenience of using a multi-turn linear potentiometer. The potentiometer is connected across the given E_s, and R_s is connected to the arm. This produces a certain non-linearity, because of the loading presented by R_s. This effect may be reduced or eliminated by using:

(a) A low resistance potentiometer, at the cost of poorer resolution and a greater load upon E_s.

(b) A cam-corrected potentiometer.

(c) A variable reference supply with low output impedance, a good though complex solution.

Adjustment of R_o

R_o may be adjusted by varying R_c, R or R_s', as shown by equation (4). R is ruled out, because it will affect E_o grossly, leaving R_c and R_s'. It is most convenient to use R_s', since R_c is usually a very small resistance.

If linear control of R_o is desired, however, R_c variation must be used. The problem of a very low resistance potentiometer can be avoided by shunting R_c with a potentiometer of more convenient value, and returning R_s' to the arm of the potentiometer.

Drift

Equation (7) gives the output drift resulting from a given drift e_p at the amplifier input. From this point of view it is clearly desirable to reduce R_o/R_c and E_{oc}/E_s. The first ratio can be reduced by increasing R_c, which tends to impair the intrinsic regulation of the degenerative circuit. It is quite practical, however, to bring R_o/R_c below unity, which is sufficient because of the unity term in equation (7).

The behaviour of E_{oc}/E_s depends upon the method of varying E_{oc}. If R_s is constant and E_s is varied from zero to $E_{s(max)}$, it may be seen from equation (6) that E_{oc}/E_s is constant, and has the value $E_{oc(max)}/E_{s(max)}$. If E_s is constant and R_s is varied, then E_{oc}/E_s will vary from zero to $E_{oc(max)}/E_{s(max)}$.

Thus it is preferable to vary R_s, from the point of view of drift, which is not surprising, since this permits the full E_s to be used at all output settings. However, in transistor supplies, where $E_{oc(max)}/E_{s(max)}$ is often not much greater than unity, the advantage is not great, again because of the unity term in equation (7).

Temperature Compensation of R_o

In the typical application shown in Fig. 2, the negative output resistance of the supply is used to cancel the external resistance r_o, resulting in zero output resistance at the ultimate load R_L. If r_o has a substantial temperature coefficient, it is desirable to introduce a compensating coefficient into the power supply, to make the r_o cancellation independent of temperature. This temperature compensation need not be made with high precision, since it is usually a second order effect. It can often be ignored entirely, in fact, although as will be seen, it can be provided for easily. To investigate this possibility, the ultimate load voltage E_L may be found from equation (5), omitting the drift term:

$$E_L = (E_s/R_s)\,R + R_o\left((E_s/R_s) + I_L\right) - r_o I_L \quad \ldots\ldots (8)$$

Because of the $R_o\,E_s/R_s$ term, it is not correct to simply give R_o the same temperature cofficient as r_o. Instead, the coefficient of R_o should be such as to make $dE_L/dT = 0$. For a given load R_L, I_L will be constant if this temperature compensation is successful—thus constant I_L may be assumed as a condition. Equation (4) shows that a temperature coefficient may be introduced into R_o through R_c, R_s' or R. It will first be assumed that this is done through R_c or R_s'—therefore the first term in equation (8) is constant. Thus:

$$dE_L/dT = \frac{\partial E_L}{\partial R_o}\frac{dR_o}{dT} + \frac{\partial E_L}{\partial r_o}\frac{dr_o}{dT} = 0 \quad \ldots (9)$$

$$R_o = r_o \quad \ldots\ldots\ldots\ldots (10)$$

where:

$$\partial E_L/\partial R_o = (E_s/R_s) + I_L \quad \ldots\ldots\ldots\ldots (11)$$

$$\partial E_L/\partial r_o = -I_L \quad \ldots\ldots\ldots\ldots (12)$$

$$\alpha_{Ro} = (1/R_o)\,dR_o/dT \quad \ldots\ldots\ldots (13)$$

$$\alpha_{ro} = (1/r_o)\,dr_o/dT \quad \ldots\ldots\ldots\ldots (14)$$

Substituting equations (10) to (14) into equation (9), the following condition for α_{Ro} is found:

$$\alpha_{Ro}/\alpha_{ro} = \frac{I_L}{I_L + (E_s/R_s)} \quad \ldots\ldots (15)$$

Unfortunately the condition specified by equation (15) is a function of I_L. However, there are two cases for which

Fig. 2. Negative R_o of supply used to cancel external resistance

it may be satisfied. In the first case $E_s/R_s \ll I_L$, giving $\alpha R_o = \alpha_{ro}$. For the second case, equation (15) is transformed by observing that $I_L = E_{oc}/R_L$, where E_{oc} is given by equation (6), resulting in:

$$\alpha_{Ro}/\alpha_{ro} = \frac{1}{1 + R_L/(R + R_o)} \quad \ldots\ldots\ldots (16)$$

It is clear that equation (16) can be satisfied if R_L is constant—this is the second possibility. In this case it is fortunate that the R_o term in equation (16) is insignificant, since it gives the possibility of adjusting α_{Ro} independently of R_o.

In the analysis above it was assumed that α_{Ro} is introduced through R_c or R_s', not through R. A similar analysis for the contrary case gives the following value for α_{Ro}:

$$\alpha_{Ro}/\alpha_{ro} = \frac{R_o}{R_L + R_o} \quad \ldots\ldots\ldots\ldots (17)$$

This condition can only be satisfied for constant R_L, and even then, the temperature coefficient adjustment will depend strongly upon R_o. This form of compensation is therefore clearly undesirable.

Producing the Required Temperature Coefficient

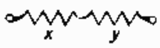
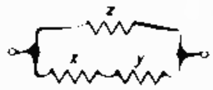
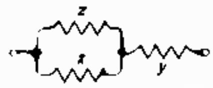
It has been shown that a certain value of α_{Ro} should be used, as given by the equivalent equations (15) and (16), and that this temperature coefficient should originate in R_c or R_s'. There are the same parameters available to control R_o, as discussed previously, with R_s' preferred for the control of R_o because of the inconveniently small magnitude of R_c. On the other hand this small magnitude is no disadvantage in controlling the temperature coefficient, because the adjustment is fixed, not continuously variable. This is of course assuming that the required coefficient is well determined, as it is when compensating copper leads.

From equation (4) it may be noted that α_{Ro} will be numerically equal to α_{Rc} or $-\alpha_{Rs'}$, whichever is used. Since copper or other metal wires, all of which have a positive temperature coefficient, are the most convenient materials for producing the desired effect, there is another reason for introducing temperature coefficient via R_c, when a positive coefficient is desired. In the typical case, therefore, R is a fixed resistor with zero temperature coefficient, R_s' is variable for adjustment of R_o, and has zero temperature coefficient, and R_c is fixed, and has a coefficient of value α_{Ro}. Output voltage is controlled by varying E_s or R_s, as already discussed.

The magnitude of α_{Ro} is $\leqslant \alpha_{ro}$, as shown by equation (15) or (16). If the equal sign holds, the required α_{Ro} can be produced by winding R_c with the same wire as r_o—usually copper. If $\alpha_{Ro} < \alpha_{ro}$, it may be possible to select a material of the required temperature coefficient, but it is usually more convenient to use a combination of high and zero coefficient resistors. The high temperature coefficient material can be the same as that composing r_o, since the addition of zero coefficient resistors will always reduce the net temperature coefficient. Table 1 presents analysis and design equations for series and series-parallel circuits, in which one resistor (x) has a temperature coefficient α_x and the others have zero coefficient. The series circuit is the simplest, but has the disadvantage that x must be wound to a definite resistance. The two series-parallel circuits give an extra degree of freedom, which may permit a given temperature-sensitive resistor x (possibly a radio frequency choke, or other available copper coil) to be used. There is nothing to choose between the two series-parallel circuits, as far as the permissible range of x is concerned, because it is the same in both cases. However, one form may be more convenient in a given case, because the zero temperature coefficient resistors required are more readily available.

TABLE 1

Terminal Resistance and Temperature Coefficient of Resistor Combinations

CIRCUIT	*A*	*B*	*C*
Schematic Temperature coefficient of $x=\alpha_x$ y, z=0			
ANALYSIS EQUATIONS			
Terminal Resistance R_t	$x + y$	$\frac{(x+y)z}{x+y+z}$	$y + \frac{xz}{x+z}$
Terminal Temperature coefficient α_t	$\left(\frac{x}{R_t}\right)\alpha_x$	$\left(\frac{x}{x+y}\right)^2\left(\frac{R_t}{x}\right)\alpha_x$	$\left(\frac{z}{x+z}\right)^2\left(\frac{x}{R_t}\right)\alpha_x$
DESIGN EQUATIONS			
x	$\left(\frac{\alpha_t}{\alpha_x}\right) R_t$	x may be within the range: $\left(\frac{\alpha_x}{\alpha_t}\right) R_t \geqslant x \geqslant \left(\frac{\alpha_t}{\alpha_x}\right) R_t$	
y	$R_t - x$	$\sqrt{\left[x R_t\left(\frac{\alpha_x}{\alpha_t}\right)\right]} - x$	$R_t - \sqrt{\left[x R_t\left(\frac{\alpha_t}{\alpha_x}\right)\right]}$
z		$\frac{R_t}{1 - \sqrt{\left[\frac{R_t}{x}\left(\frac{\alpha_t}{\alpha_x}\right)\right]}}$	$\frac{x}{\sqrt{\left[\frac{x}{R_t}\left(\frac{\alpha_x}{\alpha_t}\right)\right]} - 1}$

Example

A negative output resistance supply is to be designed with the following characteristics:

E_o = zero to 10V, with linear control by 10 turn potentiometer.

$r_o = 13\Omega$, copper.

Reference voltage = 7·5V.

$R_L = 1\,800\Omega$.

From the discussion given, it is clear that variation of E_s must be used. A potentiometer across the 7·5V reference voltage gives E_s = zero to 7·5V. If R is chosen as $1\,000\Omega$, R_s must be 750Ω, in accordance with equation (6). Since this 750Ω loads the output of the potentiometer, a low resistance potentiometer (100Ω) is used. The maximum output resistance of the potentiometer is 25Ω, at mid-position, giving a maximum departure from linearity of 3·3 per cent. If R_c is chosen as 13Ω, equation (4) shows that R_s' must be $1\,000\Omega$. To permit adjustment of R_o, to cancel the external r_o, R_s' is made up of a fixed 750Ω resistor in series with a 500Ω potentiometer, giving a range of 10 to 17Ω for R_o. As discussed, the necessary temperature coefficient$=\alpha_{Ro}$ is introduced into R_c. Equation (16) gives $\alpha_{Ro}/\alpha_{ro}=\alpha_{Rc}/\alpha_{ro}=0{\cdot}357$. In Fig. 3 α_t is the terminal temperature coefficient of the various networks, while R_t is the terminal resistance. In the present case $\alpha_t=\alpha_{Rc}$ and $R_t=R_c=13\Omega$. If the temperature sensitive resistor x has the same coefficient as r_o (both copper in this case), then $\alpha_x=\alpha_{ro}$. Therefore $\alpha_t/\alpha_x=\alpha_{Ro}/\alpha_{ro}=0{\cdot}357$.

For the simple series circuit (*A*, Table 1), $x = 4{\cdot}64\Omega$, $y = 8{\cdot}36\Omega$. Since a copper coil of the required value ($4{\cdot}64\Omega$) may not be readily available, it will be noted that circuits *B* and *C* permit x to be in the range 4·64 to $36{\cdot}4\Omega$. Choosing $x = 8\Omega$, then for circuit *B*:

$$y = 9{\cdot}1\Omega,\ z = 54{\cdot}4\Omega$$

For circuit *C*:

$$y = 6{\cdot}9\Omega,\ z = 25{\cdot}6\Omega$$

These values need not be accurate, since a continuous adjustment is provided for R_o, while the temperature compensation is only a correction of a correction. This circuit has been constructed, and worked as specified. In particular, the negative output resistance was independent of output voltage setting.

The Determination of the Impedance Matrix of a Complicated Linear Network

By W. A. Crabbe*, M.Sc.

The Kron method of obtaining the impedance matrix of a linear network is reviewed. This method is then applied to a simple three pole and finally to a two-stage amplifier with bridge feedback. The results for certain relatively complicated network configurations are simplified by using the impedance matrix. The Kron method of obtaining this matrix gives a formalized approach resulting in fewer errors.

(Voir page 775 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 782)

THE application of matrix algebra to the analysis of three pole linear networks is well advanced. In particular, the matrices of many networks can be obtained by the judicious choice of the matrix type for the simpler three poles contained therein.

However, there are other circuit arrangements, some relatively complicated, which do not fit into the above classes. K. G. Nichols[1] analysed one of these circuits using the admittance matrix. This article proposes a solution using the impedance matrix which is derived using the

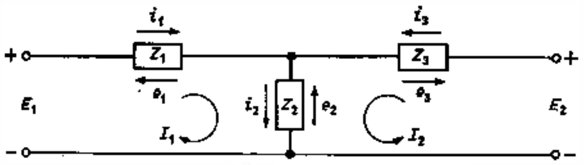


Fig. 1. Circuit for analysis

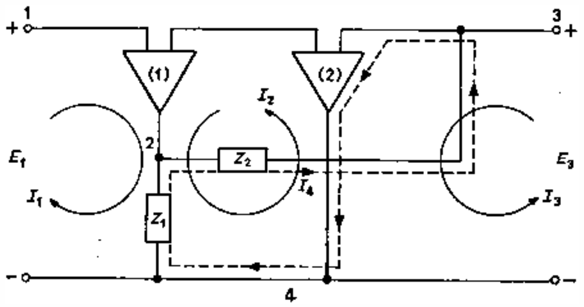


Fig. 2. Two stage amplifier with bridge feedback

method of Gabriel Kron[2]. It is suggested that the method of Kron be used to solve relatively complicated networks when the resultant matrix may be simpler than that obtained by other matrix techniques.

Kron Method of Obtaining the Impedance Matrix

(1) Write down the impedance matrix of the 'primitive' system including any mutual impedances using only the unconnected elements of the circuit $[Z']$.

(2) Find the 'connexion matrix' C which transforms the currents of the primitive system into the desired mesh currents $[i] = [C]\,[I]$.

(3) Find C_t (the transpose of C) which transforms the voltages of the primitive system into the desired applied voltages. $[E] = [C_t]\,[e]$

Thus $[e] = [Z']\,[i]$
$= [Z']\,[C]\,[I]$

and

$$[E] = [C_t]\,[Z']\,[C]\,[I]$$

or

$$[Z] = [C_t]\,[Z']\,[C]$$

(consult Kron[2] for proof).

Example (Fig. 1)

$$[Z'] = \begin{bmatrix} Z_1 & & \\ & Z_2 & \\ & & Z_3 \end{bmatrix}$$

$$\begin{bmatrix} e_1 \\ e_2 \\ e_3 \end{bmatrix} = \begin{bmatrix} Z_1 & & \\ & Z_2 & \\ & & Z_3 \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix}$$

$$\begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} = \begin{bmatrix} 1 & \\ 1 & 1 \\ & 1 \end{bmatrix} \begin{bmatrix} I_1 \\ \\ I_2 \end{bmatrix}$$

$$\therefore\ C = \begin{bmatrix} 1 & \\ 1 & 1 \\ & 1 \end{bmatrix}$$

$$\text{and } C_t = \begin{bmatrix} 1 & 1 & \\ & 1 & 1 \end{bmatrix}$$

Evaluating

$$[Z']\,[C] = \begin{bmatrix} Z_1 & \\ Z_2 & Z_2 \\ & Z_3 \end{bmatrix}$$

$$\therefore\ [C_t]\,[Z']\,[C] = \begin{bmatrix} Z_1 + Z_2 & Z_2 \\ Z_2 & Z_2 + Z_3 \end{bmatrix} = [Z]$$

$$\begin{bmatrix} \begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \end{bmatrix} = \begin{bmatrix} Z_1 + Z_2 & Z_2 \\ Z_2 & Z_2 + Z_3 \end{bmatrix} \begin{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \end{bmatrix}$$

The assumed direction of i_1, i_2 and i_3 is immaterial in this example. However, attention must be paid to the assumed current directions in the primitive circuit when mutual or transfer impedances are involved.

Three-Pole Impedance Matrix of a Two-Stage Amplifier with Bridge Feedback (Fig. 2)

In the following analysis, $\begin{bmatrix} Z_{i1} & Z_{r1} \\ Z_{f1} & Z_{o1} \end{bmatrix}$ is the impedance matrix of the first stage relative to terminal 2, and $\begin{bmatrix} Z_{i2} & Z_{r2} \\ Z_{f2} & Z_{o2} \end{bmatrix}$ is the impedance matrix of the second stage relative to terminal 4.

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$$[Z'] = \begin{bmatrix} Z_{i1} & Z_{r1} & & & & \\ Z_{f1} & Z_{o1} & & & & \\ & & Z_1 & & & \\ & & & Z_2 & & \\ & & & & Z_{i2} & Z_{r2} \\ & & & & Z_{f2} & Z_{o2} \end{bmatrix}$$

$$[C] = \begin{bmatrix} 1 & & & \\ & 1 & & \\ 1 & 1 & & -1 \\ & & & 1 \\ & -1 & & \\ & & 1 & 1 \end{bmatrix}$$

$$[C_t] = \begin{bmatrix} 1 & & 1 & & & \\ & 1 & 1 & & -1 & \\ & & & & & 1 \\ & & -1 & 1 & & 1 \end{bmatrix}$$

$$[Z'C] = \begin{bmatrix} Z_{i1} & Z_{r1} & & \\ Z_{f1} & Z_{o1} & & \\ Z_1 & Z_1 & & -Z_1 \\ & & & Z_2 \\ & -Z_{i2} & Z_{r2} & Z_{r2} \\ & -Z_{f2} & Z_{o2} & Z_{o2} \end{bmatrix}$$

$$[C_t]\,[Z']\,[C] = [Z] = \begin{bmatrix} Z_{i1}+Z_1 & Z_{r1}+Z_1 & & -Z_1 \\ Z_{f1}+Z_1 & Z_{o1}+Z_1+Z_{i2} & -Z_{r2} & -Z_1-Z_{r2} \\ & -Z_{f2} & Z_{o2} & Z_{o2} \\ -Z_1 & -Z_1-Z_{f2} & Z_{o2} & Z_1+Z_2+Z_{o2} \end{bmatrix}$$

It is possible, of course, to write down the above matrix by inspection of the circuit. However, the Kron method provides a systematic approach leading to fewer mistakes.

Since one is usually interested in the circuit performance when the external terminals (1 and 3 in this case) are excited, then the matrix is rearranged so that the meshes having no applied voltages are written down last.

e.g.

$$\begin{bmatrix} E_1 \\ E_3 \\ E_2=0 \\ E_4=0 \end{bmatrix} = \begin{bmatrix} Z_{i1}+Z_1 & Z_{r1}+Z_1 & & -Z_1 \\ & -Z_{f2} & Z_{o2} & Z_{o2} \\ Z_{f1}+Z_1 & Z_{o1}+Z_1+Z_{i2} & -Z_{r2} & -Z_1-Z_{r2} \\ -Z_1 & -Z_1-Z_{f2} & Z_{o2} & Z_1+Z_2+Z_{o2} \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \\ I_2 \\ I_4 \end{bmatrix}$$

$$= \begin{bmatrix} A & B \\ D & C \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \\ I_2 \\ I_4 \end{bmatrix}$$

where $A = \begin{bmatrix} Z_{i1}+Z_1 & Z_{r1}+Z_1 \\ & -Z_{f2} \end{bmatrix}$ etc.

It can be shown that[1]

$$\begin{bmatrix} E_1 \\ E_3 \end{bmatrix} = [A - BC^{-1}D] \begin{bmatrix} I_1 \\ I_3 \end{bmatrix}$$

$$= \frac{1}{|C|} \begin{bmatrix} \mathbf{Z_{11}} & Z_{13} \\ Z_{31} & Z_{33} \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \end{bmatrix}$$

where

$Z_{11} = Z_1 Z_{f1}(Z_{o2}-Z_{r2}) - Z_{r2}Z_2(Z_{i1}+Z_1) - Z_1 Z_{o2} Z_{i1}$

$Z_{13} = Z_1 Z_{r1}(Z_{o2}-Z_{r2}) - Z_{r2}Z_2(Z_{r1}+Z_1) - Z_1 Z_{o2}(Z_{o1}+Z_{i2}) + Z_1 Z_{r2} Z_{1}$

$Z_{31} = -Z_{f1}Z_{o2}(Z_1+Z_2) - Z_1 Z_2 Z_{o2}$

$Z_{33} = Z_{i2}Z_{r2}(Z_1+Z_2) - Z_{o2}(Z_1 Z_{o1} + Z_1 Z_{i2} + Z_2 Z_{o1} + Z_1 Z_2 + Z_2 Z_{i2}$

Conclusions

The Kron method of obtaining the impedance matri has been presented. By way of illustration, a simpl example was given.

Finally, a more complicated network was analysed an the resultant matrix reduced so that only the excitin terminals were involved.

The relative simplicity of the results shows that th impedance matrix may be preferred to other types o matrices for certain network configurations involvin 'series' elements.

Lastly, the author recommends the Kron method o obtaining the impedance matrix for complicated network because of the formalized approach and reduced chanc of errors.

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2. KRON, G. Tensors for Circuits. (Dover Publications, Inc).

A New Mass Spectrometer

The photograph shows the new analytical mass spectrometer known as type MS10c2 which has been developed by the Scientific Apparatus Department of AEI Electronics Ltd.

The spectrometer which provides for a resolution up to mass 400 is in the form of a complete console unit with built-in heated inlet system, vacuum pumps, recorder and control desk.

The basis of the spectrometer is a high resolution analyser with a high power electromagnet, the field of which can be set at any value up to 9 kilogauss, and which can be swept at six speeds from 100 to 3 000 seconds for a full scan.

Short News Items

Automatic alarm scanning and data logging equipment is being supplied to the Steel Company of Wales by Westinghouse Automation, a division of the Westinghouse Brake and Signal Co. Ltd. The equipment has been designed in co-operation with the staff of the power plant department of the Steel Company of Wales.

The power station has five turbo-alternators comprising three G.E.C. mixed pressure condensing sets, each rated at 10MVA and generating at 11kV, a G.E.C. back pressure set rated at 2 250kVA and generating at 3·3kV, and a G.E.C. back pressure set rated at 3 125 kVA and generating at 11kV. The other machine installed at present is a Sulzer Bros. turbo-blower rated at 10MW.

Initially the data logger will record 47 bearing temperatures, 3 vacuum levels, 10 steam flows, and 9 steam temperatures ranging between 350°F and 850°F. Provision has been made to log a total of 120 points to handle a possible future installation of another three turbo-blowers.

All quantities will be printed out hourly or on demand by an automatic I.B.M. Sealeotric 735 printer; if an alarm condition arises the particular quantity is printed out automatically, with a time check, by an Addo-X strip printer. The quantity continues to be printed out in red on each 2½ minute scan to show the trend of the fault condition until cleared. When the quantity has returned within limits it is printed out once only in black.

A ceramic material, known as Berylia, has been developed by Consolidated Berylium Ltd, an associate company of the Imperial Smelting Corporation of Bristol, and has found considerable application in the nuclear field as a moderator and reflector for power reactors.

This material is now finding electrical and electronic applications due to its behaviour as a good insulator and dielectric material.

It has also a low power factor over a wide frequency range and it maintains its characteristics at high temperatures. Its additional characteristics of high thermal conductivity, equal to that of aluminium makes berylia a useful material where heat dissipation and high temperature insulation are required such as heat sinks for transistors.

The Ministry of Aviation has placed with Elliott-Automation Ltd the contract for a fully automatic computerized message switching system of advanced design for the main U.K. station of the world-wide Aeronautical Fixed Telegraph Network at Croydon.

The A.T.F.N. system handles air traffic control messages on a world-wide basis. The Croydon centre, which handles over 800 messages per hour at peak times, has sixty-five channels, connecting it to other international centres and to subsidiary switching centres, air traffic control centres and airports throughout the United Kingdom and Eire.

The new system, which is fully automatic and incorporates two NCR Elliott 4100 computers, will re-route and distribute messages entirely automatically at the rate of sixty-seven words per minute over one hundred and twenty-seven channels or, alternatively, at the rate of one hundred words per minute over ninety-five channels. This capacity allows for a very substantial expansion of workload in the coming years.

'Crimped Joints for General Purpose Aircraft Electrical Cables', British Standard 2 G178, has now been published in revised form. The principal changes in the standard are the introduction of provision for insulated terminations, high temperature joints, and pin and socket contacts; and the inclusion of a wider range of cables including equipment wires.

Design requirements and tests are specified for the crimping of insulated and non insulated terminations to general purpose aircraft electrical cables, or equipment wires, with conductors of copper, copper alloy, aluminium or aluminium alloy.

Copies of BS 2 G178 may be obtained from the BSI Sales Branch, 2, Park Street, London, W.1. Price 7/6 each. (Postage 9d will be charged extra to non-subscribers).

Frequency control of the Droitwich 200kc/s transmitter has been further improved resulting in a long-term stability which is now within ±5 parts in 10^{10}. Due to the use of automatic frequency correction the excursion from nominal does not usually exceed 1 part in 10^{10}.

In 1945 the BBC inaugurated high-precision frequency control of its 200kc/s transmission of the Light Programme from Droitwich. The long-term frequency stability, which was then within 1 part in 10^7, was considerably improved in January 1963 and since then has been been maintained within 5 parts in 10^9.

With the recent increase in broadcasting hours of the Light Programme the Droitwich 200kc/s transmission is now available for 21 hours daily from 0400 to 0100 G.M.T.

Daily comparisons between the frequency of the Droitwich transmission and the National Standard of Frequency are made at the National Physical Laboratory at Teddington and the results are available by application to: The Director, The National Physical Laboratory, Teddington, Middlesex.

Ekco Electronics Ltd (Instrumentation Division) is to supply the complete nuclear instrumentation system including the control console, solid-state amplifiers and the safety system for the reactor physics research facility at the Berkeley Nuclear Laboratories of the Central Electricity Generating Board. First deliveries of equipment are due at the end of this year.

The BBC-2 transmitting station at Sutton Coldfield, near Birmingham, which was brought into service with a temporary aerial and reduced power on 6 December 1964, started transmission using the permanent aerial and increased power on 4 October. This should considerably extend the service area and make BBC-2 available to some 3½ million people in the Midlands in an area which includes most of Warwickshire, much of Leicestershire and Staffordshire and parts of Derbyshire, Shropshire and Worcestershire.

A high speed half duplex data transmission terminal ordered from G.E.C. Electronics by I.C.T. Atlas Computing Services, Manchester, will work on-line into its Manchester Atlas Computer over the G.P.O. public switched telephone network, using the G.P.O. Datel Set 1A. This will ultimately enable the scope of the Service to be expanded by catering for subscribers wishing to gain even greater advantage from the power of the Atlas computer by using it in real-time.

This order follows extensive trials carried out over a private wire between Newman Street and the I.C.T. Atlas at Manchester University. Successful completion of these trials has further demonstrated the versatility of G.E.C. Electronic's DT10A series of data links. An on-line duplex system is already working with the University of London Atlas Computing Service and more than 50 terminals for off-line systems have been supplied for a wide variety of applications.

The Government of St. Kitts has agreed to enter into a 20-year agreement under which Cable and Wireless (West Indies) Ltd will finance the expansion of the Government's existing telephone system and the improvement of the v.h.f. links between St. Kitts and Nevis, Anguilla and St. Maarten.

The company will act as consultants to the Government in connexion with the planning of the initial expansion of the telephone system and will advise the Government as to its operation and subsequent expansion.

The Company will also undertake the technical maintenance of the telecommunications equipment and power plant of the v.h.f. links.

St. Kitts is one of the islands which will benefit from the Company's £4 400 000 Eastern Caribbean expansion project. It is connected by a line-of-sight link to Antigua, whose multi-channel tropospheric scatter link with Barbados is due to go into operation this month. The whole project will be completed next year with the laying of the 80-channel telephone cable between Tortola and Bermuda, and the linking by scatter of Tortola and Antigua.

Welwyn Electric Ltd has become the first resistor manufacturer to gain full Qualification Approval to the new DEF 5115-2, Style RHF3, with its 'W' series of vitreous enamelled wirewound resistors.

DEF 5115-2 is the most stringent specification in existence for fixed high temperature resistors and supersedes the earlier DEF 5111-1.

Welwyn Electric Ltd had already received approval to DEF 5115 for metal oxide resistors.

The Ministry of Aviation has placed a contract with the Digital Systems Department of Ferranti Ltd., to supply a radar simulator system for use by the Air Traffic Control Evaluation Unit at Hurn Airport. This contract is part of a larger system that will be used for the evaluation of new air traffic control techniques and procedures, and for training of air traffic control officers.

The system consists of the Ferranti Hermes computer with silicon logic and a 2μsec, 8 192 word core store, together with extra logic and circuits which provide realistic primary and secondary radar video outputs representing up to 40 aircraft as seen from four independent radar sites. The radar site positions are program-controlled.

The system will be supplied complete with programmes to perform the simulation functions and for the preparation of exercises. The programmes will be written so as to make exercise preparation a simple procedure. An on-line Flexowriter will be provided to permit parameter changes and other supervisory actions during exercises. Diagnostic programmes will also be supplied to facilitate rapid fault-finding when necessary.

The tracks of the simulated aircraft can be either pre-programmed, or manually controlled by operators at six Aircraft Control Positions (a.c.p.) or by a mixture of both. Each a.c.p. has a Marconi tabular display for presentation of aircraft data, and a Ferranti-developed special keyboard for data entry. The use of digital equipment for radar simulation has many advantages, including accuracy, flexibility and ease of expansion. The flexibility is largely due to the ease of changing stored programmes for the computer, permitting rapid revision of the simulated a.t.c. situation between exercises.

Glasgow Corporation's Educational Television Service is now in operation and Glasgow has thus become the first city in the world to link all its schools by closed-circuit television to a central studio transmitting lessons every day during term time.

The studios and the city's 310 schools are linked by more than 100 miles of a relay system which was designed and installed in only 11 months by British Relay Ltd.

The transmitting equipment for this service is situated in the studio building in the centre of the city. Radiating from this central station are seven main cable 'feeds' comprising some 63 miles of cable laid in specially excavated trenches and 40 miles laid in disused tramway ducts. In addition to the educational channels, the network can carry the two existing BBC programmes and STV.

The extreme flexibility of the system will allow a two-way link with other educational establishments such as Glasgow University and Strathclyde University, the Royal Scottish Academy of Dramatic Art, and the Notre Dame College of Education and Jordanhill College of Education, which can both receive transmissions and produce and feed their own programmes into the network.

A revision of BS 2136—'Fixed metallized-paper dielectric capacitors for direct current' has been published. This standard covers fixed capacitors having self healing properties, a rated voltage not exceeding 6300V and containing a dielectric of impregnated paper and thin deposited metal electrodes, which are intended for use in telecommunications and similar electronic equipment.

The standard establishes uniform requirements for judging the electrical, mechanical and climatic properties of the capacitors and describes suitable testing procedures. It also classifies capacitors according to their ability to withstand conditions specified in BS 2011 (Basic method for the climatic and durability testing of components for telecommunications and allied electronic equipment). A further section, dealing with dimensions of capacitors, is still under consideration.

The revised standard is identical in technical context and layout with IEC Publication 166, with very few exceptions, and an indication of the differences is given in the standard itself.

Copies of B.S. 2136 may be obtained from the BSI Sales Branch, 2, Park Street, London, W.1. Price 17/6d each. (Postage 1/- will be charged extra to non-subscribers).

English Electric Automation announces that KOVO, the Czechoslovakian import agency, has placed orders for three computer systems with the English Electric Group worth more than £1M. These are for two LEO 360's and one KDF7, which together make the largest order Czechoslovakia has ever placed for computers. These were won for Britain against international competition.

Under the first contract, signed with English Electric-Leo-Marconi Computers, a LEO 360 computer will be supplied for extensive wagon control, stock control, payroll and other management functions, for the Czechoslovakian State Railways. This computer will be installed by the end of this year.

The second contract, signed with English Electric's Metal Industries Division, is for two computers, a LEO 360 and a KDF7, which will control the entire production of Czechoslovakia's largest steelworks, the Nova Hut Klementa Gottwalda at Ostrava-Kuncice.

This steelworks order was placed after a six-man technical team from Czechoslovakia had studied the outstanding success of the four-computer dynamic production control system installed by English Electric at Tube Investments' Park Gate Iron & Steel Works, which has been in full operation for more than a year.

BBC television trade test transmissions for BBC-1 and BBC-2 now start at 09.00.

The trade test transmission takes place daily (except Sundays). Subject to programme commitments and any alterations that may be necessary because of engineering maintenance, the schedule is now as follows:

		Vision	*Sound*
BBC-1	09·00-09·25	Test Card	Recorded music
	09·25-09·29	Test Card	440c/s tone
	09·25-09·30	Test Card	No sound

This half-hour sequence is then repeated until 13.00 and during any programme intervals between 14.00 and 17.00.

		Vision	*Sound*
BBC-2	09·00-09·04	Test Card	440c/s tone
	09·04 09·05	Test Card	No sound
	09·05-09·30	Test Card	Recorded music

This half-hour sequence is then repeated until 13.00 and from 14.00 to 18.00.

New manufacturing methods under development at the Resistor Division of the new Plessey Components Group should help to transform thin film circuit technology from a laboratory technique to production practice.

Fully automated pilot plant units now under test are capable of performing 48 manufacturing operations in a total cycle time of 13½ minutes, handling 75 circuit modules simultaneously. The equipment can be expanded to triple this capacity and a full scale mass production plant is expected to be in operation in less than 18 months. A semi-skilled operator is required only to load, start and finally unload each batch, all other operations being performed in sequence automatically.

Methods are already well established for attaching flip chip transistors to the conducting paths, while research in the deposition of photoconductive materials holds promise of means of producing precision potentiometers combining unlimited life with complete absence of noise.

The Independent Television Authority has placed a contract with EMI Electronics Ltd for the supply and installation of u.h.f. aerial and feeder systems on existing masts at the ITA transmitting stations at Durris, near Aberdeen and Dover.

Both aerials will have a 30ft aperture and will be mounted on a steel column at the top of the masts. Each will be enclosed in a 6ft diameter cylinder of glass-fibre-reinforced plastics for protection against the weather and to facilitate all-the-year-round maintenance.

Durris aerial will have an omni-directional radiation pattern and will be capable of handling four u.h.f. programmes. Dover aerial will have a directional pattern designed to avoid power wastage over the sea and interference with continental stations. The aerial will occupy only 15ft of space and will be able to transmit two programmes. Space will be left for a second aerial to handle two further programmes.

Initially the aerials will be used by the British Broadcasting Corporation and will radiate BBC 2 programmes.

Scheduled date for completion of the contract, which is worth £88 000, is Autumn 1966.

Elliott-Automation has provided the complete instrumentation system for a new large-scale, coal-dust explosion gallery at the Safety in Mines Research Establishments Field Laboratories at Buxton.

The contract, awarded to Elliott's Magnetic Tape Systems Divisions by the Ministry of Power, includes the supply of instruments to measure various phenomena associated with coal-dust explosions and of equipment for the continuous recording on magnetic tape of a number of parameters including pressure, flame-presence, flame-speed and wind-speed.

The new gallery, one of the largest of its type in the world, will be used to conduct research into the prevention of coal-dust explosions in mines under full-scale operating conditions. A coal-dust explosion usually starts with the accidental ignition of a pocket of fire-damp. The blast from this explosion raises coal-dust from the surface of the roadway and its flame ignites the dust cloud, causing a self-propagating explosion.

The main gallery, which is 8ft high, 9ft wide and 1 200ft long, has been designed to withstand pressures up to 200 lb/in². The measuring instruments have been installed in explosion-proof boxes in the gallery walls and are connected by cables to the magnetic tape instrumentation systems housed in a control building 300ft away from the main gallery.

The recording console is the centre of the tape-recording system. Information stored on the tape can be played back immediately after an experiment, if necessary, or retained indefinitely. Recording speed is 30in/sec. There are two tape decks, each recording 16 channels on 1in tape (one channel on each tape is used for timing). The record of an experiment can be reproduced through meters, displayed on an oscilloscope, or fed to a pen recorder. It could also, after processing, be fed directly to a digital computer.

The British Aircraft Industry's Aircraft Research Association has ordered a British designed and built NCR Elliott 4100 computer for advanced research into high speed flight, up to six times the speed of sound, at its Research Laboratories at Bedford, where the existing Mach 1 and Mach 4 wind tunnels are being augmented by the new Mach 6 tunnel which is nearing completion.

The extremely high velocities in the two smaller tunnels are obtained by compressing air into steel containers and releasing it into the wind tunnel under controlled conditions. The tests last only a few seconds, during which time the aerodynamic properties of experimental shapes are measured and recorded under varying conditions and airspeeds. Tests are often continued over several days during which a mass of complex data is recorded which has to be subjected to detailed analysis and examination before the final results are achieved. This work will now be largely taken over by the computer which will provide the results in a fraction of the time formerly taken. The computer will not only present the information in the form of statistical tables but will also provide graphical displays on the digital plotter which is being supplied with the system.

The computer will also be used on a wide range of theoretical and practical work carried out in the laboratories, including the design of models for experiments in the wind tunnels.

The 4100 computer, which will be delivered later this year, will include a 4120 central processor, a 6 microsecond store, a paper tape station, a control typewriter and data-disk handlers as well as the digital plotter.

The X-Ray Analysis Group of the Institute of Physics and the Physical Society is arranging a Conference on "Molecular interactions and the crystallography of ceramics" to be held at the University of Nottingham from 14 to 15 April, 1966. The conference will be divided into two main sessions dealing respectively with the crystallography of ceramics and molecular interactions. Invited speakers will include Professor H. F. W. Taylor (University of Aberdeen), Dr. B. T. M. Willis (A.E.R.E. Harwell), Mr. H. M. Richardson (British Ceramic Research Association, Stoke-on-Trent) for the first session and Professer C. A. Coulson (University of Oxford), Professor R. Mason (University of Sheffield) and Dr. C. W. Bunn (The Royal Institution) for the second session. It is also proposed to hold a third session which will be devoted entirely to contributed papers.

Short contributions of not more than 15 minutes presentation time are welcomed and outlines of about 300 words (preferably typed on quarto paper in double spacing) should be sent to the Conference Technical Secretary, Dr. S. C. Wallwork, Department of Chemistry, University of Nottingham, University Park, Nottingham, before 21 January, 1966. A leaflet (Ref ACB78) giving guidance on the preparation and submission of outlines of talks is available from the Institute and Society.

Advance registration for attendance at this meeting will be necessary and further details and application forms will be available in January, 1966 from the Meetings Officer, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

An international electronics congress organized by the German Association of Electro-Technicians (Verband Deutscher Elektrotechniker) in conjunction with the Deutsche Messe-und Ausstellungs-AG, will be held during the 1966 Hanover Fair, which takes place from 30 April to 8 May. The general themes of this congress, to be held on 4 to 6 May in the Congress Hall on the Fairgrounds, will be "Components and Applications" and "Electronics for Air- and Spacecraft".

The number of experimental pilot-tone stereophonic transmissions from Wrotham on 91·3Mc/s has been increased from two to three each week. These will be included in the Music Programme transmissions at the following times:

Monday	2.30 to 3.00 p.m.
Thursday	11.00 to 11.30 a.m.
Friday	2.30 to 3.00 p.m.

The Ministry of Aviation has placed with Plessey Radar Ltd a contract worth £¾M for the supply of a number of HF 200 heightfinders. The installations will be part of the United Kingdom multi-million pound radar scheme linking civil and military national air traffic control.

The scheme involves the integration of equipment from most of Britain's major electronic companies.

By the use of one standard heightfinder, with high data rate and increased reliability, much will be done to reduce the logistic problems arising from the maintenance of a variety of equipment. This latest HF 200 order replaces Type FPS 6 equipment.

LETTERS TO THE EDITOR

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

Transistor Circuits

DEAR SIR,—In a former note (Letter to the Editor, May 1965) I gave the h parameters for two arrangements of transistors.

As these formulas are not mentioned in classical books which deal with transistors circuits, although they are interesting for studying the properties of these configurations, I intend to establish the formulas for the most important circuits.

For a qualitative study of these devices, readers are referred to classical books.

Let h_{ij} be the hybrid parameters of a transistor connected with common-collector, then h_{ij}' and h_{ij}'' are the common-collector of VT_1 and VT_2.

Darlington's Circuit in the Common-Collector connexion

The device of Fig. 1 is easy to study because it is obtained by two three-terminal networks connected in cascade.

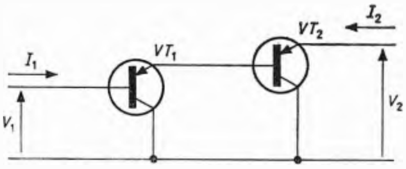


Fig. 1. Common collector connexion

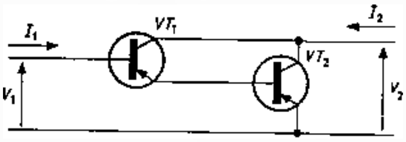


Fig. 2. Common emitter connexion

In such a case, if we define the a matrix by

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

It is known that the a matrix of the circuit is given by

$$[a] = \begin{bmatrix} a_{11}'a_{11}'' - a_{12}'a_{21}'' & a_{11}'a_{12}'' - a_{12}'a_{22}'' \\ a_{21}'a_{11}'' - a_{22}'a_{21}'' & a_{21}'a_{12}'' - a_{22}'a_{22}'' \end{bmatrix}$$

$$= \frac{1}{h_{21}'h_{21}''} \begin{bmatrix} \Delta h' \Delta h'' + h_{11}' h_{22}'' & -(\Delta h' h_{11}'' + h_{11}') \\ h_{22}'' + h_{22}' \Delta h'' & -(1 + h_{22}' h_{11}'') \end{bmatrix}$$

because of the relations

$$a_{11} = \frac{h_{12}h_{21} - h_{22}h_{11}}{h_{21}}; \; a_{21} = -(h_{22}/h_{21}); \; a_{12} = (h_{11}/h_{21}); \; a_{22} = (1/h_{21})$$

Besides, the hybrid matrix is given by

$$[h] = \begin{bmatrix} \dfrac{a_{12}}{a_{22}} & \dfrac{\Delta a}{a_{22}} \\ \dfrac{1}{a_{22}} & -\dfrac{a_{21}}{a_{22}} \end{bmatrix}$$

where

$$\Delta a = \Delta a' \Delta a'' = -\frac{h_{12}' h_{12}''}{h_{21}' h_{21}''}$$

Finally it is found

$$[h] = \begin{bmatrix} \dfrac{\Delta h' h_{11}'' + h_{11}'}{1 + h_{11}'' h_{22}'} & \dfrac{h_{12}' h_{12}''}{1 + h_{11}'' h_{22}'} \\ -\dfrac{h_{21}'h_{21}''}{1 + h_{11}'' h_{22}'} & \dfrac{h_{22}'' + h_{22}' \Delta h''}{1 + h_{11}'' h_{22}'} \end{bmatrix}$$

Darlington's Circuit in the Common-Emitter Connexion

If in Fig. 2 one exchanges the common-terminal with the output, the common-collector connexion is obtained. One is then able to write the h matrix of this device, because the common-emitter matrix is given by

$$[h_e] = \begin{bmatrix} h_{11c} & 1 - h_{12c} \\ -(1 + h_{21c}) & h_{22c} \end{bmatrix}$$

so that

$$[h] = \begin{bmatrix} \dfrac{\Delta h' h_{11}'' + h_{11}'}{1 + h_{11}'' h_{22}'} & 1 - \dfrac{h_{12}' h_{12}''}{1 + h_{11}'' h_{22}'} \\ -1 + \dfrac{h_{21}' h_{21}''}{1 + h_{11}'' h_{22}'} & \dfrac{h_{22}'' h_{22}' \Delta h''}{1 + h_{11}'' h_{22}'} \end{bmatrix}$$

where h' and h'' are the common-collector parameters of VT_1 and VT_2.

Darlington's Circuit in the Common-Base Connexion

If in Fig. 3, one permutes the common-terminal with the output-terminal, the common-emitter connexion is obtained. The h common-base matrix is given by

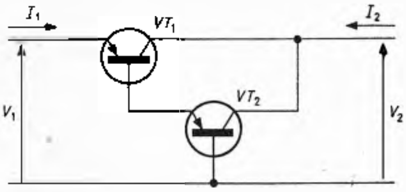


Fig. 3. Common base connexion

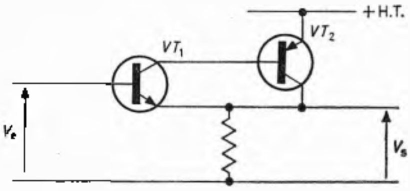


Fig. 4. Circuit with pnp and npn transistors

$$[h_b] = \frac{1}{\Delta h_c} \begin{bmatrix} h_{11c} & \Delta h_c + h_{21c} \\ (\Delta h_c - h_{12c}) & h_{22c} \end{bmatrix}$$

The calculation of Δh_c leads to the relation

$$\Delta h_c = \frac{\Delta h' \Delta h'' + h_{11}' h_{22}''}{1 + h_{11}'' h_{22}'}$$

The h matrix of the common-base connexion is then given by:

$$[h] = \begin{bmatrix} \dfrac{\Delta h' h_{11}'' + h_{11}'}{\Delta h' \Delta h'' + h_{11}' h_{22}''} & 1 - \dfrac{h_{21}' h_{21}''}{\Delta h' \Delta h'' + h_{11}' h_{22}''} \\ -1 + \dfrac{h_{12}' h_{12}''}{\Delta h' \Delta h'' + h_{11}' h_{22}''} & \dfrac{h_{22}'' + h_{22}' \Delta h''}{\Delta h' \Delta h'' + h_{11}' h_{22}''} \end{bmatrix}$$

TABLE 1

PARAMETER	COMMON-BASE		COMMON-EMITTER		COMMON-COLLECTOR		FIG. 6
	TRANSISTOR	DARLINGTON	TRANSISTOR	DARLINGTON	TRANSISTOR	DARLINGTON	
$h_{11}(\Omega)$	40	40·5	2000	93000	2000	93000	1860
h_{12}	4.10^{-4}	53.10^{-4}	16.10^{-4}	9.10^{-2}	1	0·9095	0·9986
h_{21}	−0·98	−0·9996	49	2270	−50	−2271	−2220
$h_{22}(\Omega^{-1})$	10^{-6}	10^{-6}	50.10^{-6}	$2{\cdot}34.10^{-3}$	50.10^{-6}	$2{\cdot}34.10^{-3}$	$2{\cdot}32.10^{-3}$

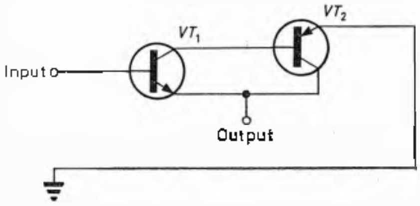


Fig. 5. Equivalent circuit of Fig. 4

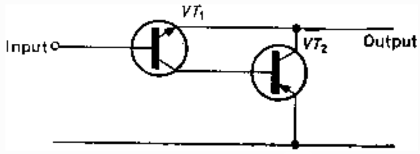


Fig. 6. Development of Fig. 5

The h matrix of this circuit will be the h matrix of the Darlington circuit in the common-emitter connexion, but the parameters h' will be the common-emitter parameters of VT_1 while the h'' will be the common-collector parameters of VT_2.

Thus one is able to calculate the hybrid parameters of the Darlington circuits from the common-collector hybrid parameters of the transistors.

Association of pnp and npn transistors

The device of Fig. 4 was considered in the note already mentioned.

By developing the incremental equivalent circuits, one obtains successively Fig. 5 and Fig. 6.

$$[h] = \begin{bmatrix} \dfrac{h_{11}' + \Delta h' h_{11}''}{1 + h_{11}'' h_{22}'} & 1 - \dfrac{h_{12}' h_{12}''}{1 + h_{11}'' h_{22}'} \\ \dfrac{h_{21}' h_{21}''}{1 + h_{11}'' h_{22}'} - 1 & \dfrac{h_{22}'' + \Delta h'' h_{22}'}{1 + h_{11}'' h_{22}'} \end{bmatrix}$$

The device of Fig. 6 is close to the Darlington circuit in the common-emitter connexion, but for transistor VT_1 emitter and collector have been exchanged.

Numerical data is shown in Table 1.

Yours faithfully,

J. AIDLER,
University of Brussels.

A Transistorized Electrometer-Amplifier

DEAR SIR,—For the measurement of small currents, in nuclear reactor instrumentation, usually down to 10^{-12}A, amplifiers with electrometer valves are normally applied. An electrometer valve being directly heated is difficult to use as a differential amplifier which is attractive with respect to low drift. Another disadvantage is microphonics. Using a m.o.s. dual transistor input stage these drawbacks can be avoided. An amplifier with this type of transistor has been developed here and may be of interest for readers working with small current measurements. The amplifier described is intended for use with feed-

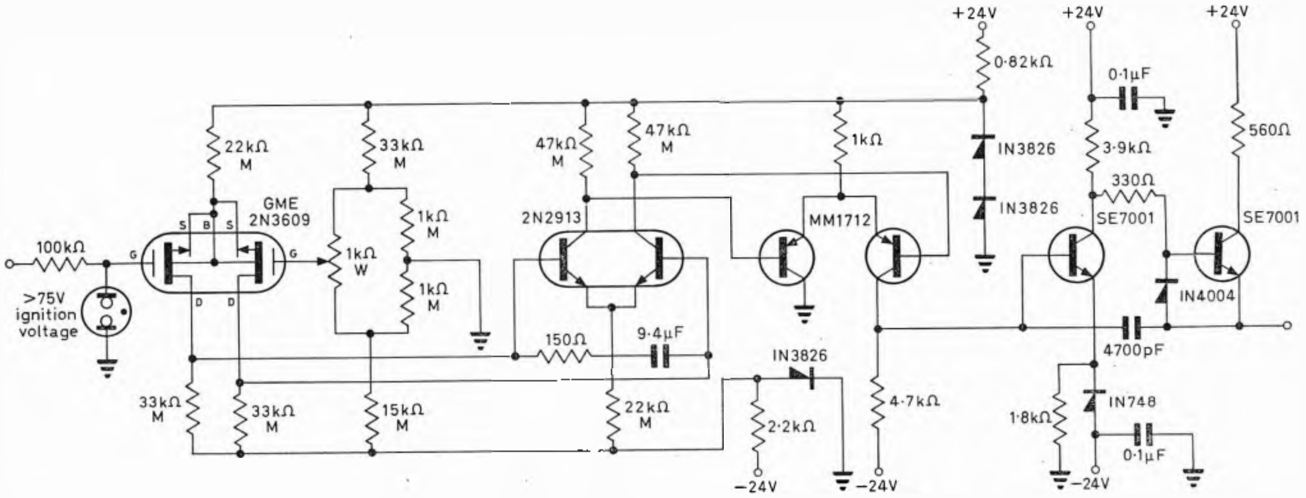
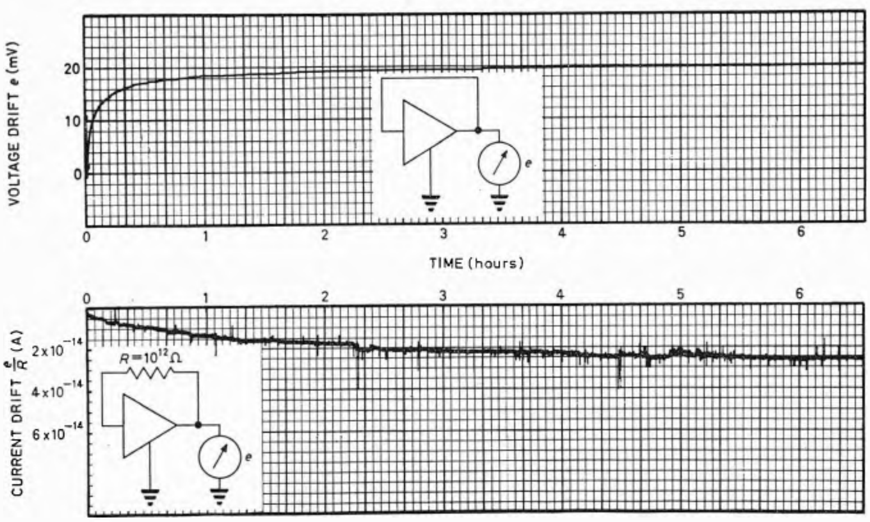


Fig. 1. The electrometer-amplifier

Fig. 2. Typical drift curves



back like an analogue operational amplifier as a linear or logarithmic amplifier, etc., and may be used for current measurements down to approximately 10^{-14}A. Measured data is given in Table 1.

The following comments apply to the circuit given in Fig. 1. The neon bulb used for gate protection must be shielded against light[1]. An error current of 10^{-12}A due to sunlight was measured with an unscreened lamp. The output circuit[2] incorporates transistors rated to V_{CEO} = 150V with a view to a possible voltage range of $\pm$50V.

The m.o.s. input transistor (as well as an electrometer valve) has initial drift when supply voltage is applied. For the amplifier described here, typical drift curves obtained after long time (hours) supply interruption are shown in Fig. 2.

TABLE 1
Amplifier Data. Drift and noise referred to input

24 hours current drift	2.10^{-15}A
24 hours voltage drift	1mV
Temperature drift 20°C to 50°C ...	54µV/°C
Supply voltage sensitivity	3mV/V
Open loop d.c. gain	5 000 times
0dB gain frequency	100kc/s
Output voltage range at 20mA load	±10V
Noise peak-to-peak	1mV

Similar drift is described as resulting from charge migration[3] at the transistor surface due to the large field strength induced in the gate dielectric, when a signal is applied and before the enhancement of conduction begins. After short interruptions (seconds) the initial drift measured was small compared to the drift illustrated.

Yours faithfully,

K. S. Højberg,
Atomic Energy Commission,
Denmark.

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An Exponential Pulse Generator

Dear Sir,—The need often arises, in the calibration and testing of nucleonic equipment, to subject the equipment to a similar pulse to that generated in radiation detectors. Although rectangular pulse generators are invariably used with success in most cases, a few arise where this shape of pulse is not suitable; e.g. where delay line pulse shaping is used, setting up of single and multi-channel pulse height analysers.

The exponential pulse generator described was designed for use in calibrating and testing equipment of the type mentioned. Another requirement was to have a continuously variable amplitude control of high accuracy and repeatability. The p.r.f. is fixed by the mains frequency.

The basis of the generator (Fig. 1) is a mercury wetted relay. This is a polarized type having an extremely fast make and break time with no contact bounce[1]. the only slight disadvantage is that the relay must be operated within 30° of the vertical. The type used has an international octal base, and is driven at mains frequency.

The voltage across RV_1 is supplied from a high performance series stabilizing circuit[2]. The input to the stabilizer can be either from two 8·1V mercury cells in series, having a capacity of 1Ah, or from a 14V d.c. mains supply having a ripple of 2mV peak-to-peak. The output of the stabilizing circuit is set to 10V and is stable to within +10mV for an input voltage change between 11V and 16V. Sk_3 and Sk_4 provide a means of monitoring this voltage during use.

The amplitude of the output pulse is set by adjustment of RV_1 which is a 500Ω 10 turn Helipot having a linearity of ±0·25 per cent and a tolerance of ±1 per cent. During one half cycle of the mains frequency the 1µF capacitor is allowed to charge, via the 470Ω series resistor, to a potential determined by the wiper of RV_1. On the other half cycle the capacitor is discharged into an attenuator network having an input resistance of 200Ω and an output resistance of 50Ω. With no attenuation switched in, the output pulse amplitude will be half the value at the wiper of RV_1.

The output pulse amplitude is continuously variable between 0V and 5V; the rise time of the pulse is about 10nsec and the decay time-constant is 200µsec. S_2 is provided to reverse the pulse polarity.

Yours faithfully,

G. Burton,
Bridgwater, Somerset.

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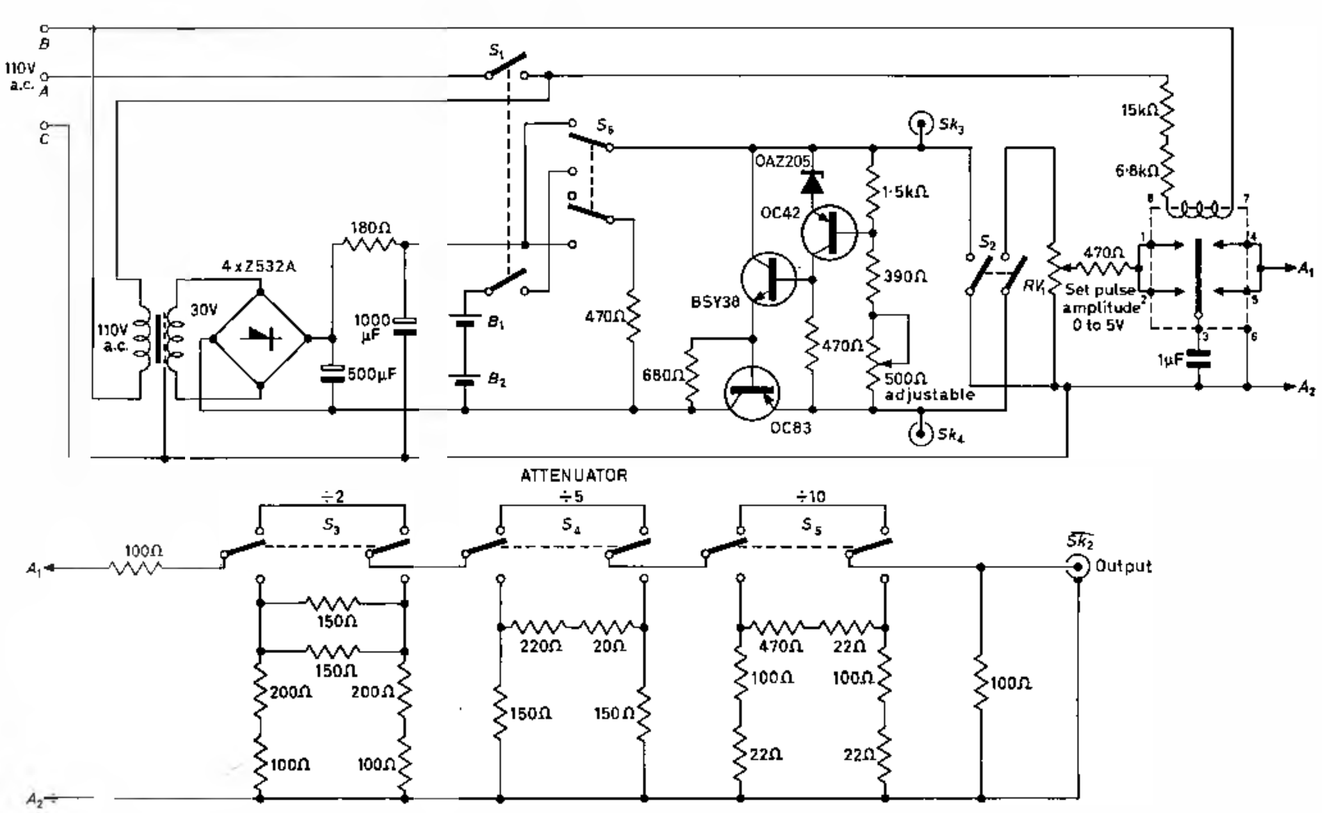
Fig. 1. The exponential pulse generator



BOOK REVIEWS

Basic Theory of Space Communications

By Frederick J. Tischer. 463 pp. Med. 8vo. D. Van Nostrand & Co. Ltd. 1965. Price 92s.

THIS is a scholarly and—to a practising engineer—somewhat academic treatment of four main themes in the theory of space communications, viz.:

(1) Electromagnetic theory of aerials.
(2) Electromagnetic theory of plasma.
(3) Wave propagation.
(4) Communication theory.

The emphasis throughout is on fundamental theory and the analytical approach. Within certain limits the treatment is thorough and well documented, with a bibliography and problems to be solved at the end of each chapter. The text is based on lectures by the author, who is Professor of Electrical Engineering at the University of North Carolina, U.S.A.

Much of the material has been published elsewhere, but the author has done a service in bringing it together in a logical and well co-ordinated manner. As such it will serve as a useful reference book for those concerned with research and advanced development in the field of space communications, and as a text for post-graduate training purposes. It is not, however, a particularly helpful book for those concerned with the more practical aspects of space communication system design and operation, or those without a good academic background in physics or electrical engineering. In many cases the solution to a problem is left in the form of an equation without a clear indication of the magnitudes of the quantities involved or of their physical significance. Some of the references are to company reports which may not be readily accessible to the general reader. In spite of these limitations the book can be recommended for specialists and those who have the appropriate background.

W. J. BRAY

Biological Effects of Magnetic Fields

Editor: Madeleine F. Barnothy. 324 pp. Med. 8vo. Plenum Press, New York. 1964. Price $16.00.

THE subject of this book recalls the early attempts to link the magic of the lodestone with the miracle of life, which led to the discredited study of animal-magnetism. Between them the 31 contributing authors should convince the most sceptical reader that some magnetobiological effects undoubtedly exist. Their hypotheses may not stand up to further research, but a serious attempt has been made to establish another respectable branch of science. Much encouragement has come from the U.S. space programmes and cancer research foundations; generous patrons who want firm answers concerning the survival of men—or their component cells—in abnormal physical environments. This book is the first to give results of these and related investigations, together with introductory chapters designed to assist newcomers to the field.

The book's 27 chapters are divided into five sections. Part 1 provides a potted explanation of magnetic theory, defines the terms and units (c.g.s.), and explores possible magnetobiological interactions at the level of organism, cell, molecule and atom. In being the work of six authors it fails to provide the unified approach necessary for such an introduction. One surprisingly meets magnetism *via* gravitation and a possible unified field theory, but reverts to Faraday's lines and iron filings after one paragraph. The subsequent mixture, including imaginary poles and quantum mechanics, is of uneven texture.

Time-varying fields are excluded because induced electrical effects swamp those depending on magnetism alone. Although direct coupling between practical steady fields and complex molecules produces changes much smaller than the normal kinetic energy, some reversible changes occur when entities possessing magnetic moments or various degrees of dia- and paramagnetism are subjected to strong fields, strong gradients or both together. The resulting alignments or displacements may be slight, but biological amplification can in time produce observable and often irreversible changes in the organism.

Parts II and III, the bulk of the book, are collections of separate papers, averaging 10 pages in each, describing in detail specific *in vivo* and *in vitro* experiments. Many details will interest only specialists working on the same problems, but general readers who skip these will notice that the conclusions reached are not always wholly consistent with each other. Fields of a few thousand oersted, with varying degrees of inhomogeneity, have retarded the growth of young mice and postponed ageing, caused the rejection of transplanted tumors, accelerated the growth of barley seedlings, inhibited bacterial growth, increased enzyme activity and reduced mutations in fruit flies. It is even claimed that a few hundred oersted can produce conditioned reflexes in fish, and alter the elctroencephalogram of rabbits. On the other hand, men have been unaffected by exposure to 20kOe, and mice have thrived after exposure to fields six times as strong.

Perhaps the most interesting, but least convincing, section of the book is Part IV., containing three chapters on the effect of weak fields and gradients, comparable to those existing in nature. The authors spurn the results of their strong-field colleagues as sledge hammer effects, and claim that the unknown sensors are saturated by fields much greater than 10Oe. Refined statistical analysis has shown that certain planarians and snails tend to move in directions related to field geometry, but the same experiments also reveal a positive correlation with the sun's position and the phases of the moon. On a much less rigorous basis, the water dowser's reflexes are related to changes in the geomagnetic gradient. Two magnetic mechanisms are proposed to explain the navigation of migrating birds. In one the flapping wings act as choppers to detect small electric potentials induced in the tissues, and in the other the relativistic geoelectric field produces signals up to 1mV across birds changing direction while in flight. The book ends with Part V containing biographical notes, an index and a 15 page bibliography that supplements the references given at the end of each chapter.

Between them, the Editor and her husband contribute nine chapters. Being physicists they emphasize the need to control or allow for all variables present during an experiment. Control experiments are essential if statistical analysis is to be of value, and they must match the actual experiments as closely as possible. For example, the pole-pieces must be simulated because these may cause cosmic ray showers, produce temperature changes, alter convection currents and reduce illumination, all of which can have biological effects. Not all the authors have taken such care over their experiments. While only those involved with magnetobiology should consider purchasing this volume, parts could be read with profit by anyone interested in the less familiar frontiers of science. Even if scepticism is not overcome, any tendency to dogmatic assertions about the impossibility of various interactions should decrease in the face of the evidence produced.

A. S. VELATE

Mathematical Analysis of Observations

By B. M. Shchigolev. 350 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 63s.

THIS is a translation, edited by H. Eagle, of a Russian textbook first published in 1960. The book is based on lectures for astronomy students at the Moscow State University. It should be of interest to scientists and technologists who are concerned with the interpretation of experimental results.

The book is divided into five parts (and contains a short bibliography of relevant Russian literature).

Part I (3 chapters, 42 pages) deals with operations with approximate numbers. An account is given of the errors incurred during addition, subtraction, multiplication, division and functional evaluation. The magnitude of the error is estimated and it is shown how this may depend upon the computational procedure employed.

Part II (3 chapters, 52 pages) is concerned with polynomial interpolation of a function of a single variable. The methods of divided and finite differences are employed for interpolation with variable and constant intervals of tabulation respectively. The interpolation formulae of Lagrange, Newton, Bessel and Stirling are included but that of Everett is omitted. There is no mention of the well-known 'throw-back' technique.

Part III (6 chapters, 108 pages) contains a standard account of probability theory. After basic concepts, the addition and multiplication theorems are introduced and the binomial distribution is derived. Subsequently the Normal, Poisson, Charlier, Pearsonian and Maxwellian distributions are described. Special attention is given to the probability of events in a large number of trials and to the joint probability distribution of two continuous variates. The Chebyshev-Markov lemma and the theorems of Bernoulli, Laplace and Lyapunov are explained.

Part IV (5 chapters, 80 pages) is entitled 'Fundamentals of the Theory of Random Measurement Errors'. Types of error and the evaluation of probable error for both unweighted and weighted measurements are described. There is a detailed explanation of the method of least squares for evaluating the coefficients in an empirical formula.

Part V (2 chapters, 51 pages) deals with the analysis of statistical material in order to obtain a qualitative criterion. Use is made of mean, median, variance and moments for univariate distributions and of linear regression and correlation analysis for bivariate distributions.

Although each part is complete in itself, the five parts are blended well into a book which is amply supplied with worked examples.

S. KIRBY

Advances in Control Systems —Volume 1

Edited by C. T. Leondes. 362 pp. Med. 8vo. Academic Press. Price 93s.

THIS book, the first in a series compiled with the intention of presenting up-to-date information in the field of automatic control, is divided into six separate reviews.

M. Aoki of the University of California discusses optimal and sub-optimal policies in control systems. Stochastic and adaptive control systems, and the realization of desired trajectories are considered. J. S. Meditch of the Aerospace Corporation presents the Pontryagin maximum principle. He applies this to the problem of specifying the optimal thrust programme for the vertical flight of a rocket. P. K. C. Wang of I.B.M. explores the extent to which the results associated with lumped parameter systems can be applied to distributed parameter systems. H. Halkin of the Bell Telephone Laboratories approaches some important concepts in the theory of optimal control from the study of systems described by difference equations. P. R. Schultz of the Aerospace Corporation considers optimal control in the presence of state vector measurement errors. Finally F. H. Kishi of T.R.W. discusses on-line computer control techniques, and applies them to the re-entry flight control of an aerospace vehicle.

The level of the reviews is aimed between that of the technical journal and the research monograph, and they should be of interest to workers developing or making use of advanced control theory.

G. D. BERGMAN

Photoelectric Materials and Devices

By S. Larach. 434 pp. Med. 8vo. D. Van Nostrand. 1965. Price 93s.

The connexions between light and electrical energy are dealt with in this book which also includes such subjects as injection-type lasers and the photoconductive, photovoltaic effects.

The book consists of the contributions of 14 individual authors each engaged in the physics, chemistry and applications of photoelectronic materials.

European Miniature Electronic Components and Assemblies Data Part II

By G. W. A. Dummer and J. Mackenzie Robertson. 1123 pp. Demy 4to. Pergamon Press. 1965. Price £12 10s.

The second part of this series covers the miniature electronic components and assemblies manufactured in France, Holland, Scandinavia and Switzerland.

As with Part I which covered the similar products in Germany and Italy, this part contains a six-language glossary of component titles and microelectronic terms.

Also included are some 250 abstracts of articles dealing with components and assemblies, together with a number of tables, charts and applications formulae.

Recent Advances in Selenium Physics

Edited by European Selenium-Tellurium Committee. 160 pp. Med. 8vo. Pergamon Press. 1965. Price 80s.

This volume contains the proceedings of a symposium on Solid and Liquid State Selenium Physics held at the Chemical Society London in June 1964 at which 13 papers were presented.

The conference was sponsored by the European Selenium-Tellurium Committee (E.S.T.C.) formed in September 1963 by the three major producers of primary selenium-tellurium in Europe.

Advances in Electronic Circuit Packaging Volume 5

Edited by L. L. Rosine. 297 pp. Crown 4to. Plenum Press. 1965. Price $15.00

The fifth volume of this series contains the 22 papers presented at the fifth International Electronic Circuit Packaging Symposium held at Boulder, Colorado, in August 1964.

This Symposium was sponsored by the University of Colorado, EDN (Electrical Design News) and Design News.

Mathematics and Computer Science in Biology and Medicine

317 pp. Med. 8vo. Her Majesty's Stationery Office. 1965. Price £3

This volume contains the proceedings of the Conference held at Oxford in July 1964 at which 26 papers were presented.

The volume also contains biographical notes of the various contributors, together with a list of delegates attending the conference which was sponsored by the Medical Research Council in association with the Health Departments.

ELECTRONIC EQUIPMENT

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

(Voir page 769 pour la traduction en français; Deutsche Übersetzung Seite 776)

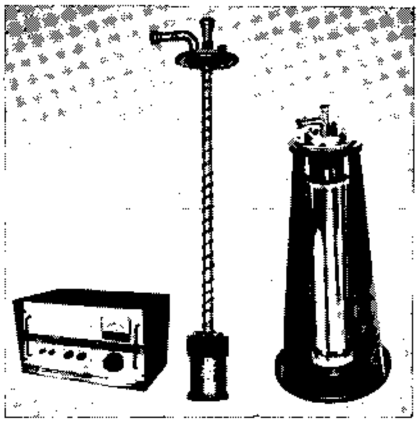
SUPERCONDUCTING SOLENOID SYSTEMS

Distributed by: Cryosystems Ltd, 40 Broadway, London, S.W.1

(Illustrated below)

A new range of superconducting solenoid systems for generating high magnetic fields—manufactured by International Research & Development Co. Ltd is available from Cryosystems Ltd.

A major advantage of these systems is that, while retaining all the essential features of more elaborate and more expensive equipment, they are simple and low-priced. The basic system, consisting of solenoid assembly, cryostat and stand, power supply, syphon and accessories, and which is centred around a 40kG solenoid having a working bore of 1·5cm, may be used in a wide variety of solid state and other research experiments.



Solenoid systems incorporating special design features can also be supplied to conform with the specific requirements of customers.

The illustration shows the power unit, the standard solenoid and cryostat is supported by a hard wood stand and the illustration shows the solenoid mounted in position.

EE 87 751 *for further details*

RESPONSE CONTROL UNIT

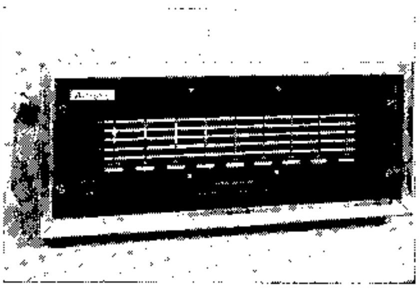
Associated Electronic Engineers Ltd, Dalston Gardens, Stanmore, Middlesex

(Illustrated above right)

This instrument enables infinite variation of the level of a.f. signals in octave steps from 40c/s to 10 240c/s. The nine separate controls each allow continuously variable adjustment of +13dB −13dB.

Designed for broadcasting studio use, it also has many applications in recording studios. Incorporated in factory p.a. systems it enables overall system response to be set for maximum intelligibility.

The standard model A1646 has input and output impedances of 600Ω at 0dBm level, and zero insertion loss. An 'in—out' switch is fitted and gain control. Overload level is +18dBm. Response is within ±2dBm 20c/s to 20kc/s: Noise—60dB.



The standard model is housed in a metal cabinet with illuminated octave controls the settings of which are displayed in graph form.

Other models are available for rack mounting and with alternative input and output arrangements.

EE 87 752 *for further details*

DUAL TRACE OSCILLOSCOPE

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

Developed primarily for field service of high-speed, solid-state computers, the type 453 combines the performance and features of a laboratory oscilloscope in an instrument only 6¾in by 10¾in by 9in, and weighing 28 lb. It uses a new Tektronix 4in c.r.t. designed to provide the high writing rate and brightness required for use under high ambient light conditions.

Dual-trace sensitivity is to 20mV/div at 50Mc/s, to 5mV/div at 40Mc/s, and the channels can be cascaded to obtain 1mV/cm sensitivity at 25Mc/s, single-trace. Signal delay allows viewing the leading edge of the triggering waveform.

Four operating modes include each channel singly, alternate or chopped electronic switching between channels, or both channels added algebraically. In the added algebraically mode, channel 2 can be inverted for differential operation.

Horizontal deflexion facilities include calibrated sweep delay for convenient measurements such as pulse-to-pulse intervals and amount of jitter on computer signals or any train of pulses, time-differences, and phase angles.

Calibrated sweep extends from 5sec/div to 0·1μsec/div, with a 10× magnifier extending the fastest sweep to 10nsec/div.

Complete, full-passband triggering facilities include automatic, single-sweep and channel 1 only. For added operating ease, a light indicates when the sweep is receiving triggers.

An X-Y display can be obtained using channel 1 for the X input. This provides 5mV/div sensitivity at 5Mc/s on the horizontal axis.

A calibrator, accurate within 1 per cent in amplitude and frequency, furnishes 1V and 0·1V 1kc/s square waves for convenient calibration. A 5mA current-probe loop is also provided.

EE 87 753 *for further details*

PHOTO-ELECTRIC TAPE READER

Great Northern Telegraph Works, 5 St. Helen's Place, London, E.C.3

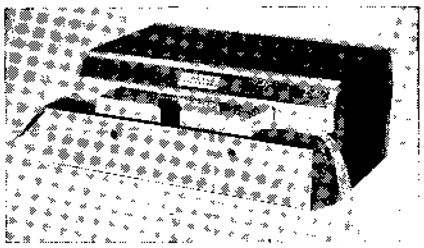
(Illustrated below)

This bi-directional photo-electric punched paper tape reader is manufactured by the Great Northern Telegraph Works of Denmark.

It has been designated the model 4101 and its main application is as a read-in unit for electronic digital computers and similar equipment requiring high-speed operation. The continuous operation speed of the instrument can be varied from 250 to 1000 characters/sec and the step-by-step operation speed available is to 500 characters/sec on the special step-by-step input and up to 1000 characters/sec with a process time (time from start of sprocket hole to new start) of the order of 60μsec, minimum. Reading is instantly adjustable for 5-, 6-, 7-, or 8-channel tape. The speed of the tape is monitored and can be read on an internal scale. The combination of the characters over the reading station is shown on the read-out panel.

The reader has been designed to make the insertion of the tape easy, as the tape latch, when open, allows the tape to be accurately positioned over the light window of the reading station. As the tape latch is lowered three fingers grip the tape keeping it in the correct position until the brake takes over when the tape latch is fully closed. Tape threading through capstans, guide rollers and the reading position is not necessary since the tape latch automatically positions the tape to feed correctly. A special light guide ensures uniform illumination of all channels in the tape. The light intensity is maintained at a constant level regardless of supply voltage fluctuation or ageing of the lamp, which is run 30-40 per cent below specified ratings to give a long lifetime, by the use of a servo amplifier.

It follows that an instrument operated



at such a high reading speed must be capable of being stopped and started very rapidly. In the model 4101 this is accomplished by powerful electromagnets. The tape can be stopped in 0·8msec and accelerated to reading speed from the stop position in approximately 2 to 3msec.

The four stage read-out amplifiers supply square output pulses with a rise and fall time of less than 2·0μsec and lengths dependant on tape speed. The sensitivity of the amplifiers can be regulated internally. Also provided are three driver amplifiers for the electromagnets and a stabilized power supply equipped with silicon rectifiers. Circuits are temperature compensated to permit operation over a 60°C ambient temperature range. A separate blower ensures good cooling of all components. The electronic input and output specification makes the tape reader compatible with most transistorized electronic equipment.

Alarm circuits give information to the associated electronic equipment in respect of open tape latch, tape out and manual control of the tape feed.

Provision is made for monitoring of most voltages in the power supply.

This instrument is offered to operate from either 220V a.c. 50c/s or 115V a.c. 60c/s and has a power consumption of approximately 150W. It can be supplied with character parity check at an extra charge, this feature being obtained by plugging two printed circuit boards into already wired connectors.

EE 87 754 *for further details*

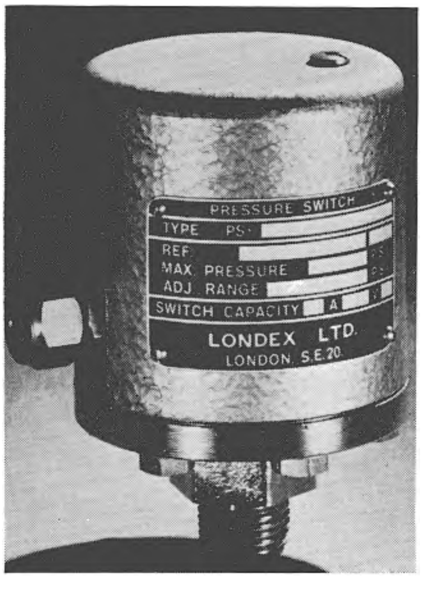
PRESSURE SWITCHES

Londex Ltd, 207 Anerley Road, London, S.E.20

(*Illustrated below*)

A new range of pressure switches with both an adjustable contact reset differential and an adjustable operating range is available from Londex Ltd, a member of the Elliot-Automation Group.

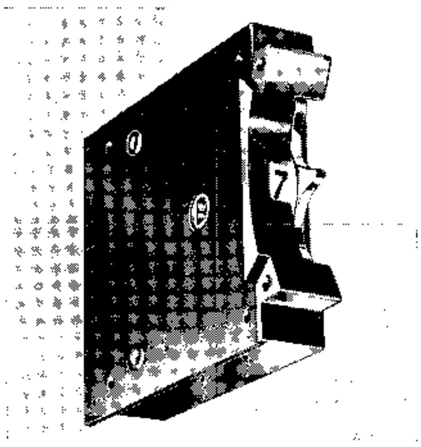
Four models of the new type PS/U switch are available, covering operating pressures of from 10 to 250 lb/in². The adjustable reset differentials are from 1 to 4 and 3 to 10 lb/in² depending on the pressure range of the particular switch.



These new pressure switches are based on a hydraulically formed seamless phosphor-bronze bellows which acts at the pressure sensitive element. Expansion or contraction of the bellows operates a snap-action single-pole changeover switch which has the following typical ratings: 5A at 440V a.c., 10A at 250V a.c., 0·2A at 220V d.c. non-inductive.

The operating characteristics of the PS/U range are illustrated by the example of a switch with an adjustable range of between 30 and 100 lb/in² which has been set to operate its contacts when pressure falls to 45 lb/in². As the reset differential of this particular device can be varied between 1 and 6 lb/in², the contacts in the minimum position will reset when pressure rises to approximately 46 lb/in². If the control is set to the maximum, the reset pressure will be approximately 51 lb/in².

EE 87 755 *for further details*



THUMBWHEEL SWITCHES

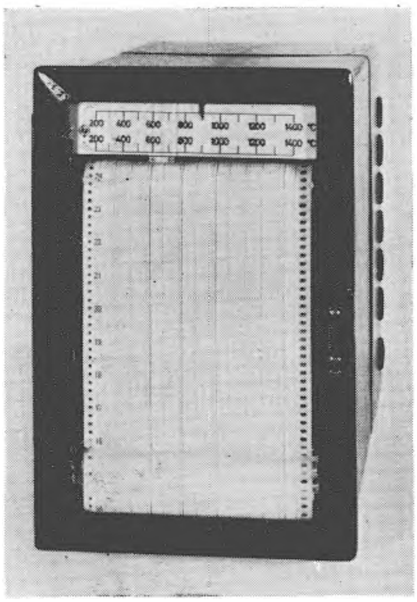
Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(*Illustrated above*)

A range of digital thumbwheel switches has been introduced by Digitizer Techniques Ltd. The switch, known as the Digiswitch, has been designed for use in applications where in-line read-out of a 'set up' number is required. Modular construction associated with a printed circuit switch board has resulted in a unit of ½in width which enables high packing densities to be achieved. Contact arrangements are normally 10 or 12 way single pole, but 5 pole 2 way or 2 pole 5 way are also available if required. The contacts have been designed to switch a.c. or d.c. of up to 50V at 100mA, but the continuous current rating can be as high as 1A. These ratings have been achieved by using a standard silver graphite contact in association with a hard gold plated printed circuit, the substrate of which is normally s.r.b.p., but fibreglass may be provided to special order.

An unusual feature of this switch is the provision of rear illumination of the characters as an optional extra. Several different voltage bulbs may be specified, typical examples being 6, 12 and 24V.

EE 87 756 *for further details*



SIX CHANNEL CHART RECORDER

Hewlett Packard, Dallas Road, Bedford

(*Illustrated above*)

Six channels of information can be plotted at the same time with the new model 7190C six channel controlling recorder. For use in industrial or scientific measurements, the model 7190C utilizes six solenoid driven ink cartridges mounted in a single assembly which virtually eliminates the moving parts typical in conventional designs. Fully transistorized, it is designed for unattended, trouble free operation.

A series of optional control features is offered for those users who need to control operations as well as produce visual records. These options include a control potentiometer which moves with the recording head and up to three limit switches for off-on control. These options may be delivered with the recorder or added at any time later.

A variety of input modules is available covering temperature spans of 100°C to 1 600°C, voltage spans from 5mV to 250V and current spans from 250mA to 5A. Other input modules are available on special order.

EE 87 757 *for further details*

MILLIVOLT AMPLIFIER AND INDICATOR

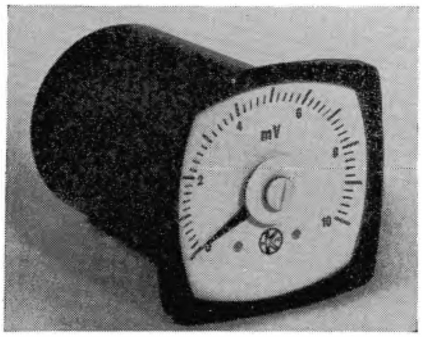
Kent Precision Electronics Ltd, Vale Road, Tonbridge Kent

(*Illustrated on page 764*)

Kent Precision Electronics Ltd has introduced a new millivolt indicating instrument as part of the range of KPE industrial solid state control instrumentation.

The instrument consists of the KPE low drift transistor direct coupled amplifier, which feeds a standard approved type indicating ammeter. The entire instrument is built within a standard 3½in panel mounting instrument case.

The amplifier is suitable for a minimum effective input range of 0 to 10mV and the input can also be adapted for connecting resistance elements in bridge



form to provide accurate indication. The instrument normally contains an accurate reference voltage. The amplifier has a current output of 0 to 1mA with very high output impedance and is suitable for feeding external remote indicators, recorders and controllers in addition to the internal indicating ammeter.

The instrument has wide applications for temperature indication, recording or control using either thermocouples (when automatic cold-junction-compensation can be provided) or resistance thermometers. It can be used with the KPE range of solid state control instruments, providing a current signal suitable for operation of a standard KPE solid state calibrated control instrument, in addition to its normal function of indication.

EE 87 758 *for further details*

COMMUNICATIONS RECEIVER

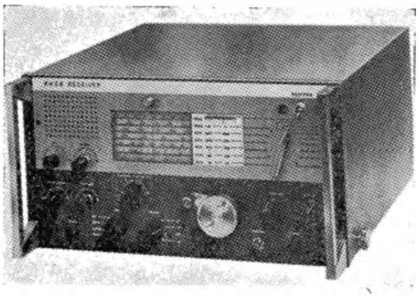
Redifon Ltd, Communications Division, London, S.W.18

(Illustrated below)

Redifon Ltd has developed a highly sensitive and stable fully transistorized communications receiver, designed for general purpose telephony, telegraphy or facsimile reception in the 13kc/s to 28Mc/s range.

The equipment, known as the Redifon R.408, which has been type approved by the G.P.O., is suitable for point-to-point services, fixed or mobile stations and for marine use. High standards of reliability are achieved by extensive use of silicon planar transistors.

The R.408 has a continuously variable bandwidth and can be used to receive on c.w., a.m., s.s.b. or any one sideband of i.s.b. (switched selection of upper or lower sideband). Full selectable a.g.c. is provided with attack and decay times to suit all services. A new type of circuit designed by Redifon gives the fast attack time and long decay time required for c.w. and s.s.b. without the disadvantage of 'paralysis' by bursts of noise or interference.



Faciilties for full s.s.b. (pilot or fully suppressed carrier) are provided and for any one sideband of an i.s.b. transmission. The bandwidth is continuously variable from 800c/s to 8kc/s for a.m. and c.w. and from 800c/s to 4kc/s for s.s.b. reception. When operating on c.w. a narrow bandwidth of 160c/s can be switched in via a crystal filter.

The control layout was designed using the results of a wide study of operating practices. A 90in film tuning scale eliminates the need for a separate logging scale and crystal calibration is provided at 100kc/s intervals.

Audio, line i.f. and a.g.c. outputs are provided and full protection is afforded against open- and short-circuits and against signals from an associated transmitter even when the R.408 is switched off.

The standard R.408 will operate from 100 to 125V or 200 to 250V, 50 to 60c/s a.c. power supplies and special models will operate from a 24V or 110V or 220V d.c. power supplies.

EE 87 759 *for further details*

PRECISION POTENTIOMETER

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

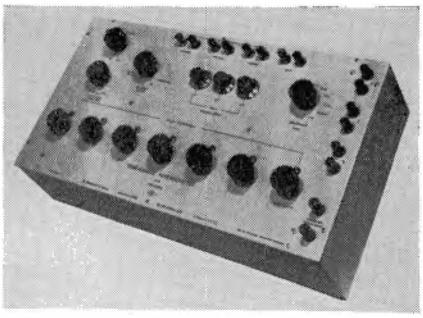
(Illustrated right)

H. Tinsley & Co. Ltd has recently added to its range of potentiometers a precision instrument designed to overcome the limitations of the conventional d.c. potentiometer. Known as the 'Stabaumatic' type 5545, it incorporates a new design principle based upon the Tinsley Automatic Current Controller, type 5388; making the instrument independent of battery stability and therefore available for immediate use. Stability is to three parts in a million. For work of the highest precision an independent temperature controlled standard cell battery may be used.

A constant resistance-potential circuit having no switch contacts is employed and there are seven measuring dials in conductance form with binary steps. Selection of any one of three measuring circuits is by means of a 5-position switch. The fourth position on the switch selects the internal standard cell and the fifth position converts the instrument into a resistance measuring bridge having six ranges. A three-dial standard cell setter, connected across the potential circuit, allows the potentiometer to be standardized to one part in 10^6 of the e.m.f. of the internal or an external standard cell.

The usual feedback resistor standard in the automatic current controller circuit is replaced by the switched conductance units of the potentiometer, the total current passing through these units in series with the potential circuit. The potential provided is proportional to the current, which is varied by operation of the conductances by means of the measuring dials. Voltage drop across those conductances which are switched into the measuring circuit is maintained constant and equal to the e.m.f. of a standard cell by means of the automatic current controller. When the conductances are operated, the current in the constant-resistance potential circuit is varied in such a way that the conductance dial readings are equal to the voltage across the circuit. This voltage is balanced against an unknown voltage by means of the galvanometer detector in the usual manner. Those conductances which are switched out of the measuring circuit are connected so that they still form part of the total current path.

All conductances are switched in parallel by the dials to produce decade values although the number of coils is only four per dial; making adjustment easier and eliminating resistance errors as the contacts are in relatively high resistance circuits and are in parallel. There are no thermal errors introduced by switching in the potential circuit which is a constant resistance of 100Ω on the top range and 10Ω on the low range. This provides a thermal free potentiometer for high precision low voltage thermocouple work.



Standardization of the instrument is effected by setting the measuring dials to the known value of a standard cell inside the instrument and adjusting the standard cell setter to give the correct balance. When done, standardization is maintained by one of the internal cells balancing the controller. Two internal control cells are provided. One acts as a buffer cell during preliminary balance when out-of-balance currents may temporarily upset the cell balance. The buffer cell and second, or working, control cell are selected by the sensitivity control switch, so that the working cell is not brought into use until a close balance has been obtained. A third standard cell is used for standardization. Temperature compensation is effected automatically for ambient temperature variations of 19°C to 24°C. Provided the ambient temperature remains within this range, the standardization need only be checked against the internal cell once or twice in a day.

Two measurement ranges are provided. These are 0 to 2·11111V in steps of 1μV and 0 to 211·111mV in steps of 0·1μV. Over these ranges the instrument has a calibrated accuracy of ±0·001 per cent of reading +1μV and a long term limit of error of ±0·003 per cent of reading +1μV. The range of the instrument as a bridge is from 0·01Ω to 21MΩ and the limit of error is 0·005 per cent or 0·0005Ω over the

0·01Ω to 21kΩ range, then 0·01 per cent up to 211kΩ, and 0·1 per cent up to 2·1MΩ and 1 per cent to 21MΩ. For this application the automatic current controller is not required, but it is an essential part of the circuit when the instrument is operated as a potentiometer.

EE 87 760 *for further details*

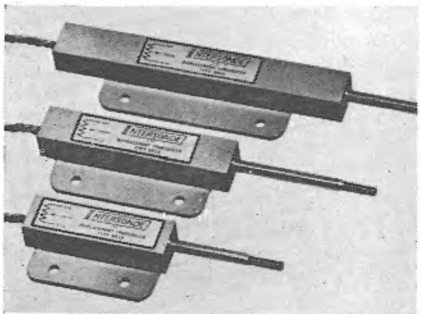
DISPLACEMENT TRANSDUCERS

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustrated below)

The type DR68 potentiometric displacement transducers manufactured by Intersonde Ltd are available with stroke lengths of ½, 1 and 2in and are based on a precision wirewound resistance element. In these devices rectilinear displacement of the sensing shaft drives a noble-metal wiper over the resistance element which is normally energized with a constant voltage. The wiper voltage is directly proportional to the applied displacement and at full rated displacement the wiper voltage is substantially equal to the excitation voltage. With a suitable excitation the transducer output is sufficient to operate many types of indicators and recorders without amplification being necessary.

The transducers provide a reliable and low-priced means of sensing displacement and offer an accuracy of better than 0·5 per cent of full range and a resolution of up to 0·001in. The operating temperature limits extend from −20°C to +120°C and the life expectancy of the transducers exceeds 1 000 000 operations.



EE 87 761 *for further details*

TAPE COMPARATOR

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

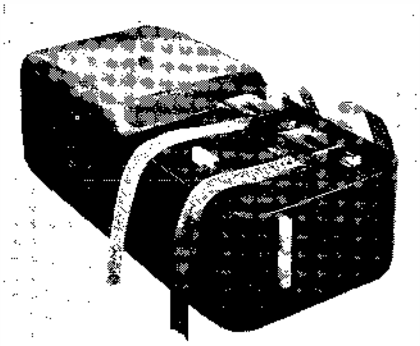
(Illustrated above right)

The 'Aptec' tape comparator model AT9102, is one of the smallest, fastest and most self-contained units of its kind available. It incorporates a modified version of the Danish G.N.T. twin-head tape reader, and has facilities for comparing 5, 6, 7 or 8 level punched tapes in any code regime.

Designed as a means of checking the accuracy of copy tapes produced from originals, it has a maximum speed of 30 codes per second.

The 'Aptec' comparator drives the original and copy tapes to be compared from a common motor shaft through two separately controlled reading heads. When the coding sensed by both heads agrees exactly, the drives to the two tape transports are simultaneously energized and both tapes 'step' on to the next code position. If dissimilar codes are sensed the drives are suppressed and both tapes stop. After checking discrepancies, tapes may be realigned on the next identical code by means of either 'advance' or 'single shot' buttons. When the codes agree again the comparator will proceed automatically. Both reader heads are fitted with 'tight-tape' and 'tape-out' interlocks which cause the machine to revert to 'stop' in the case of tape faults. Similar interlocks also operate if either reading head is left open. The 'tape-out' interlocks will also detect damage to the sprocket holes of the tapes.

A feature of the 'Aptec' comparator is the ease of loading, which is accomplished by simply laying the tapes on the sprockets in the open gate of each head. The gates are then closed when the machine is ready to operate. The



'single shot' and 'advance' buttons enable very rapid alignment of the tapes prior to comparison.

The machine will operate through heavy splices, including p.v.c. joints, and is a self-contained unit measuring 18·5in by 8in by 7in.

EE 87 762 *for further details*

STABILIZED POWER SUPPLIES

Kingshill Electronic Products Ltd, Torrens Works, Torrens Street, London, E.C.1

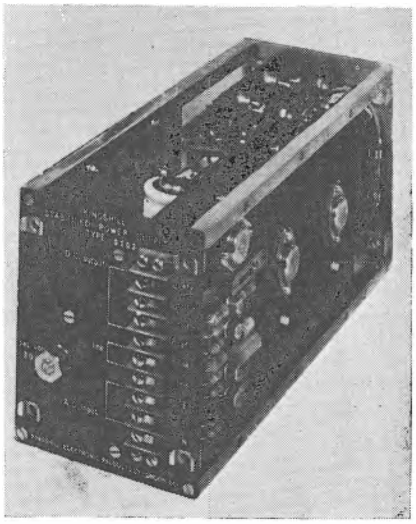
(Illustrated above right)

The G series stabilized power supplies is a range of high performance modular units covering 0 to 30V up to a maximum current of 10A.

Low cost has been achieved by rationalizing both electrical and mechanical design. Best quality components are used throughout, e.g. C-core transformers and high stability resistors.

The clean mechanical construction allows these units to be mounted from any face with the additional advantage that they are not restricted to the customary upright mounting position, it being permitted to stand them on end. These features are especially advantageous as a variety of mounting methods are thus available when replacing existing units of different manufacture.

Two basic panel sizes are used, 5¼in by 4in and 5¼in by 5½in, the length of the units ranging from 2¾in to 15in, giving the utmost flexibility in stacking behind a standard 19in panel. The module sizes range from 5⅛in by 4in by 2¾in for a 0 to 30V ½A or 0 to 15V 1A rating, to 5⅛in by 5½ in. by 15in for a



0 to 15V 10A rating.

Each unit has a self-resetting electronic current limiter which may be preset over a wide range, and a preset output voltage control allowing a swing of about 5V. Remote sensing is standard and the positive rail is brought out to a link for current monitoring or alarm.

Regulation is better than 0·1 per cent, stabilization ratio 1 000 : 1, ripple 1mV, mains tolerance ±10 per cent, maximum ambient temperature 45°C.

EE 87 763 *for further details*

HEAVY DUTY RELAY

Magnetic Devices Ltd, Newmarket, Suffolk

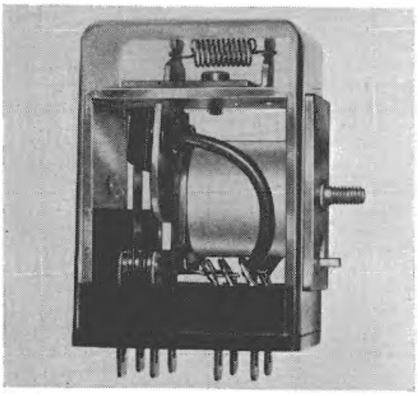
(Illustrated on page 766)

A rugged double pole, double throw relay, designed to combine maximum reliability with maximum economy, has been introduced by Magnetic Devices Ltd. Suitable for both solder connexions or 'quick connect' receptacles, all terminations are conveniently grouped together in a moulded housing, and adequate arc barriers and generous clearance between live parts make it suitable for arduous applications.

Available both for a.c. and d.c., the relays are protected by a transparent Makrolon cover and the special moulding materials used throughout give high anti-tracking properties. Mounting is by means of a single 3 B.A. stud in association with a locating spigot.

Series 140EP (a.c.) is rated at 10A and has a nominal coil power of 5·5VA, maximum working voltage of 440V, and maximum coil resistance of 10kΩ. Series 145EP (d.c.) is rated at 10A and has a nominal coil power of 1·75W, maximum working voltage of 150V, and maximum coil resistance of 14kΩ. (A series resistor can be supplied separately for series 145EP for higher voltages.)

Average operate and release speeds of the series 145EP are 25msec. The relays will operate in ambient temperatures of



up to 35°C, but can be adapted for higher temperatures if specified.

Maximum weight is 8 oz, and fine silver or elkonite contacts are available. They are also available with vacuum impregnated coils for humid or tropical conditions.

EE 87 764 *for further details*

TEMPERATURE CONTROLLERS

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

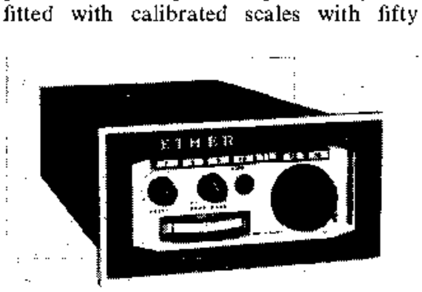
(Illustrated below)

A new series of precision thermocouple temperature controllers designated the series 18-90 has been announced by Ether Ltd. Four types are at present available.

These temperature instruments have three main sections, these being a precision set-point, a deviation amplifier and a controller. They can be operated from normal mains supplies in ambient temperatures from 0 to 50°C.

On the types 18-92 and 18-98 a seven-position push-button switch and a 10-turn precision potentiometer provide the set-point setting. Each switch position gives steps of 100° and the potentiometer divides each 100° step into a thousand divisions. This gives a settability of 0·1°. The input signal is filtered and then compared with the output signal of the set-point section. The resultant error signal is fed into the deviation amplifier, the output of which is fed to a deviation meter. On the type 18-92 the output shown on the deviation meter is fed into the control section to give a 0 to 5mA d.c. signal. On the type 18-98 the output is arranged to give firing pulses suitable for driving thyristors.

The types 18-99 and 18-92 are identical to the 18-92 and 18-98 respectively except that they are designated as differential thermocouples controllers. In place of the digital set-point they are fitted with calibrated scales with fifty



equal divisions. The standard calibration is −0·5mV to +0·5mV.

EE 87 765 *for further details*

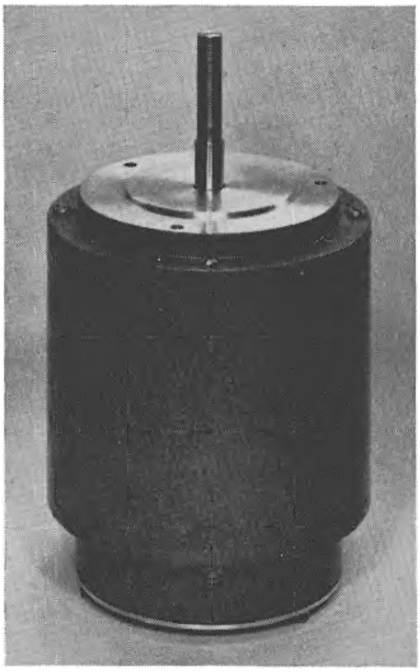
LOW-SPEED SYNCHRONOUS MOTOR

Vactric Control Equipment Ltd, Garth Road, Morden, Surrey

(Illustrated below)

Designed for applications where a slow, constant speed is required, such as office copying machine operation, gate valve regulation or machine tool indexing control, a new version of the 'Vacsyn' slow-speed synchronous motor is now entering production at the Morden factory of Vactric Control Equipment Ltd.

To facilitate compact installation in restricted areas, the overall diameter of

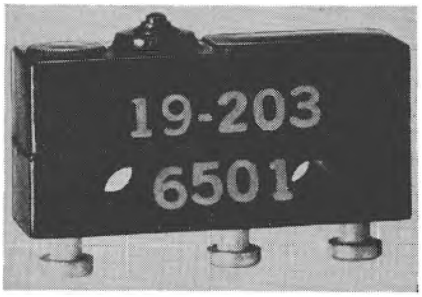


the motor has been reduced to less than 4in, but the original power output of 6·6W is retained. In its standard form the motor is wound for 115V 50c/s giving a minimum torque of 150 oz.in and versions with speeds of either 60 or 71·4rev/min are available. Other windings and speeds can be made to special order.

The low synchronous speed of the motor is attained by the use of a permanently magnetized multi-pole rotor with a stator having a balanced two-phase winding. It is designed to run from a single phase supply, the phase displacement for the second winding being provided by an external series resistor and capacitor.

No expensive gearing is necessary and the motor is self-starting and reversible. It can also be used as a stepper motor by applying a sequence of positive and negative pulses to each winding. Each change of polarity causes the shaft to advance either 1·8° or 2·14° depending on motor type, so that either 200 or 168 steps are taken to complete one revolution.

EE 87 766 *for further details*



MICROSWITCH

The Plessey Co. Ltd, Electromechanical Division, Havant, Hampshire

(Illustrated above)

A new subminiature microswitch introduced by the Electromechanical Division of the Plessey Components Group combines high sensitivity with long mechanical life.

Known as the type 19, it features a switch blade mechanism which involves only two moving parts, having a movement differential for operation of as little as ·0005in maximum, for ultra-fast and sensitive action. Mechanical life is in excess of 10 million cycles, while electrical life at the full rated load of up to 5A at 250V a.c. is 100 000 cycles. Nominal electrical life at half rated load is 250 000 cycles.

The switch is basically a single pole, double throw unit, but is also available in single throw normally open and normally closed forms. Solder and turret type terminals are offered, and the standard contact material is silver. Contacts in gold plated silver or special materials are optionally available.

Standard operating force for the type 19 microswitch is 4 to 5 oz, and a full range of actuators may be fitted. For normal applications the case, which measures only 0·25in by 0·78in by 0·36in, is moulded in phenolic resin, with diallyl material available for use up to 200°C.

EE 87 767 *for further details*

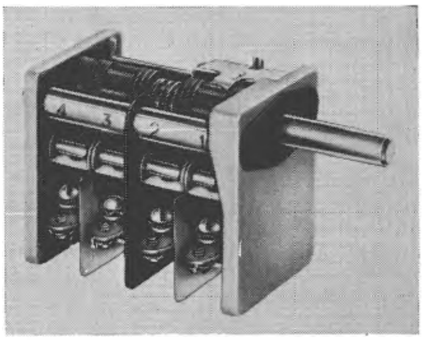
ROTARY SWITCH

Honeywell Controls Ltd, Brentford, Middlesex

(Illustrated below)

Honeywell Controls Ltd has produced a new versatile rotary switch which can replace the functions of up to eight toggle switches.

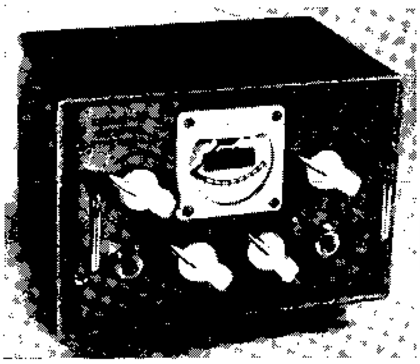
Known as the 1RM, this compact unit occupies only 1·625 by 1·45in of panel space. It is ruggedly built from Honeywell's high capacity V3 switches. These switches can be operated in any sequence



—with up to 40 000 variations—and the same unit can house switches suitable for different current levels; gold contacts being used for low energy circuits and silver contacts for 15A circuits. Operation of the switches is by a camshaft, constructed to give precision sequencing over a very long life—with either manual or motor powered operation.

One advantage associated with the new unit is that the snap action of the V3 switches eliminates the white noise troubles normally associated with rotary wafer switches. True panel sealing of the unit is achieved with an adhesive-faced gasket and 'O' ring.

EE 87 768 *for further details*



SELECTIVE AMPLIFIERS

Avon Communications and Electronics Ltd, 14-16 High Street, Christchurch, Hampshire

(Illustrated above)

Intended for a wide variety of applications in both industrial and telecommunications fields, a range of inexpensive and versatile tunable selective amplifiers has been recently introduced by Avon Communications & Electronics Ltd. They are known as the 400 series.

The amplifiers are continuously tunable over a frequency range from 270c/s to 3 700c/s in four bands. Selectivity can be varied between 5 per cent and 20 per cent and good rejection of frequencies adjacent to the pass-band is ensured by the use of two simultaneously tuned circuits giving an attenuation characteristic which approaches 120dB/octave.

Active elements are transistorized and powered by internal batteries giving a life of approximately three months. The amplifiers have a compact and rugged construction and are fitted with a 2½in meter for signal tuning purposes.

When used in telecommunications networks, the performance of the selective amplifier is further improved by employing a non-linear output stage, or 'gate', since by imposing a threshold or minimum signal level at the output a useful improvement in signal-to-noise ratio can be obtained.

Such an instrument can improve the signal-to-noise ratio of a telecommunications link by at least 20dB.

In laboratory applications, such as bridge null detection, linear output stages can be fitted and selective amplifiers can provide a substantial improvement in the accuracy of null points. Models can be provided with a built-in limiter circuit to obviate oral shock when used with bridge applications and headphone detection.

EE 87 769 *for further details*

REED UNISELECTORS

Alma Components Ltd, Park Road, Diss, Norfolk

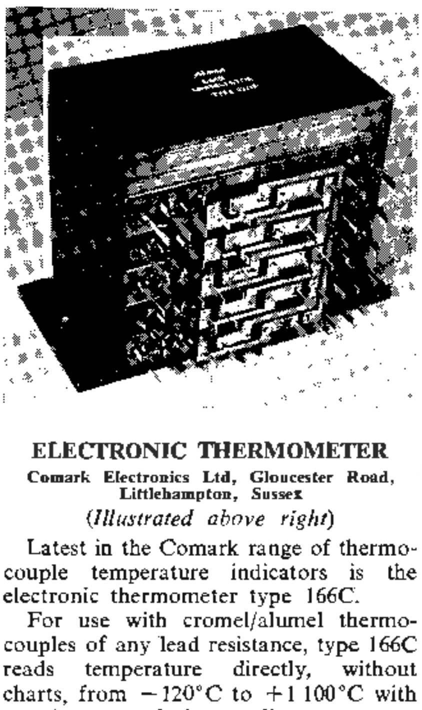
(Illustrated below)

These new uniselector reed units are small, reliable, flexible and considerably less expensive than other electronic switching devices; a 3-bank 10-way unit measures only 3½in by 3½in by 3in complete. All that is required is a 24V supply and the unit will then step forward one position when two wires from the unit are shorted together—no additional circuits are required.

Reverse operation can be provided and automatic homing to any pre-selected number can be incorporated. Facilities for connecting the units in decades with a 'carry-forward' from zero to the next unit can be built in if required. Any number of ways up to 100 can be supplied with up to six banks. Changeover or normally closed contacts can be provided instead of the standard normally open contacts.

Many advantages result from using a reed system; low power requirements (½W), high speed of operation, fully sealed contacts requiring no maintenance or adjustment coupled with long life, independent contacts instead of common wiper circuit and low contact resistance.

EE 87 770 *for further details*



ELECTRONIC THERMOMETER

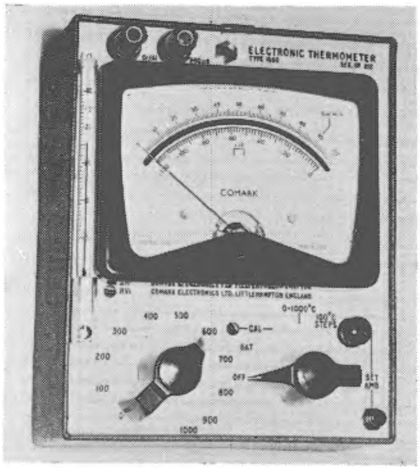
Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(Illustrated above right)

Latest in the Comark range of thermocouple temperature indicators is the electronic thermometer type 166C.

For use with cromel/alumel thermocouples of any lead resistance, type 166C reads temperature directly, without charts, from $-120°C$ to $+1\,100°C$ with one-degree resolution at all temperatures.

Auto cold-junction, solid-state chopper amplifier and long-life mercury batteries



give rapid, accurate measurements, with the minimum of controls.

The thermocouple output voltage is backed-off in 100°C steps by a precision potential divider calibrated against an internal reference cell. The meter spans the 100° intervals. A search range of 0 to 1 000°C is also provided.

Standard recorder output is 200μA.

EE 87 771 *for further details*

SEMICONDUCTOR TESTER

Distributed by: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(Illustrated below)

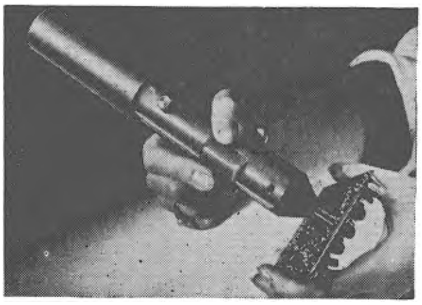
The 'Diotestor' manufactured by Sercel S.A.R.L. of France is designed specially to test semiconductors on printed circuits at the end of a production line. It detects faulty diodes and transistors without the need for unsoldering these components.

The Diotestor checks conduction and cut-off conditions in pnp/npn transistors and diodes. Test procedure is simple and requires no special knowledge.

Two models are available, one using a standard 3V dry-battery and weighing 6 oz, the other using Voltabloc accumulators (rechargeable by plugging directly into the mains) and weighing 11 oz.

Self-powered and in an unbreakable Rilsan housing, its handy shape and lightweight make the Diotester easy and convenient to use.

Two accessories are available, an adaptor with seven different spacings for testing diodes, and an extending lead with socket and crocodile clips to test unmounted semiconductors.



EE 87 772 *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.

Ordinary Meetings

Date: 4 November.
Discussion: The Selection of Topics for Research in Electrical and Electronic Engineering. (Joint with the I.E.R.E.)
Opened by: W. Jackson and L. Rotherham.

Control and Automation Division

Held at Hatfield College of Technology, College Lane, Hatfield, Herts.
Date: 8 November. Time: 6 p.m.
Discussion: Is Laboratory Work Really Necessary?
Opened by: K. R. Sturley.
Date: 9 November.
Lecture: The Design of the A.C. Servomotor.
By: D. E. Hesmondhalgh.

Electronics

Date: 1 November. Time: 9.30 a.m. and 2.15 p.m.
Colloquium: HS 303 (Early Bird) and the Post Office Earth Station at Goonhilly.
(All wishing to attend must register: forms available from the the Secretary, I.E.E.)
Date: 3 November.
Paper on: Atmospheric Radio Noise and Signals Received on Directional Aerials at High Frequencies. (Paper E4529.)
Date: 3 November.
Paper on: Characteristics of Atmospheric Radio Noise Observed at Singapore. (Paper E4736.)
Date: 5 November.
Lecture: Operational Experience with Tropospheric-scatter Systems.
By: R. W. Cannon, G. C. Rider and D. Wilkenson.
Date: 15 November. Time: 10.30 a.m. and 2.30 p.m.
Colloquium: Phase Measurement throughout the Spectrum.
(All wishing to attend must register: forms available from the Secretary, I.E.E.)
Date: 17 November. Time: 6 p.m.
Lecture: Electroacoustics.
By: E. Mayer.

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Electro-Acoustics Group

Held at: The I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 3 November. Time: 6 p.m.
Lecture: High Intensity Acoustic Noise Sources.
By: B. L. Clarkson.

Joint I.E.E./I.E.R.E: Research Committee Meeting

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 4 November. Time: 5.30 p.m.
Discussion: The Selection of Topics for Research in Electrical and Electronic Engineering.

Joint I.E.E./I.E.R.E. Medical Electronics Groups

Held at: The I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 10 November. Time: 6 p.m.
Discussion: Medical Electronics Round the World.
Opened by: D. W. Hill, R. Brennand, W. J. Perkins and H. S. Wolff.

Radar Group

Held at: The I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 17 November. Time: 6 p.m.
Lecture: The Use of Doppler Radar in Meteorological Research.
By: P. G. F. Caton.
Date: 22 and 23 November.
Conference on: U.H.F. Television.
(Joint I.E.R.E./I.E.E. Registration forms and further information from the I.E.R.E., 9 Bedford Square, London, W.C.1.)

Joint I.E.R.E./I.E.E. Computer Groups

Held at: The I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.
Date: 24 November. Time: 6 p.m.
Lecture: Analogue Circuit Techniques Using Transistors.
By: G. J. Crask and S. F. Miles.

Southern Section

Held at: Highbury Technical College, Cosham, Portsmouth
Date: 3 November. Time: 6.30 p.m.
Lecture: Satellite Communications.
By: A. E. Bailey.

Held at: Bournemouth Municipal College of Technology and Commerce.
Date: 10 November. Time: 7 p.m.
Lecture: Ultrasound in Medicine.
By: P. N. T. Wells.
Held at: Brighton College of Technology.
Date: 16 November. Time: 6.30 p.m.
Lecture: Digital Computers in Road Traffic Control.
By: H. A. Codd.
Held at: Basingstoke Technical College.
Date: 25 November. Time: 7.30 p.m.
Lecture: Difficulties in the Use of Coded Systems for Information Transfer at Machine Interfaces.
By: D. Proctor.

South Western Section

Held at: University of Bristol Engineering Laboratories.
Date: 1 November. Time: 6 p.m.
Lecture: Measurements at Sub-Audio Frequencies. (Joint meeting with the Western Centre of the I.E.E.)
By: H. Sutcliffe.
Held at: City of Plymouth College of Technology.
Date: 16 November. Time: 7 p.m.
Lecture: Television Recording.
(Joint meeting of the Western Centre of the I.E.E.)
By: P. Leggatt.
Held at: University of Bristol Engineering Laboratories.
Date: 17 November. Time: 7 p.m.
Lecture: Automatic Dough Production in a Biscuit Factory.
(Joint meeting with the Institution of Production Engineers.)
By: D. R. Pentelow and A. Wilkinson.
Held at: Bath Technical College.
Date: 28 November. Time: 7 p.m.
Lecture: Infra-red Technology and Applications.
By: V. Roberts.

South Wales Section

Held at: Welsh College of Advanced Technology, Cardiff.
Date: 10 November. Time: 6.30 p.m.
Lecture: Underwater Acoustical Engineering.
By: D. G. Tucker.

East Anglian Section

Held at: Chelmsford Technical High School, Patching Hall Lane.
Date: 9 November. Time: 6.30 p.m.
Lecture: The Applications of Integrated Circuits.
By: G. C. Padwick.
Held at: University Engineering Department, Trumpington Street, Cambridge.
Date: 25 November. Time: 8 p.m.
Lecture: Inospheric Propagation with special reference to the Effect of the Earth's Magnetic Field.
(Joint meeting with the Cambridge Electronics Section of the I.E.E.)
By: G. Millington.

Yorkshire Section

Held at: Doncaster Technical College.
Date: 4 November. Time: 7 p.m.
Lecture: Electronics in Coal Mining.
By: E. J. Ferrier.

East Midland Section

Held at: Loughborough College of Technology.
Date: 16 November. Time: 7 p.m.
Lecture: The BBC Radiophonic Workshop
(Joint meeting with the I.E.E.)
By: F. C. Brooker.

West Midland Section

Held at: Wolverhampton College of Technology.
Date: 10 November. Time: 7.15 p.m.
Lecture: London-Birmingham-Manchester Radio-Relay Link.
By: E. M. Hickin.
Held at: University of Birmingham
Date: 28 October. Time: 6.30 p.m.
Lecture: The Early History of Radio.
By: G. R. M. Garratt.

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle.
Date: 10 November. Time: 6 p.m.
Lecture: The Integrated Systems of Photography.
By: M. J. Savage.

North Western Section

Held at: Renold Building, Manchester College of Science and Technology, Sackville Street, Manchester.
Date: 18 November. Time: 7 p.m.
Lecture: Low-Noise Solid-State Amplifiers.
(Joint meeting with the I.E.E.)
By: H. N. Daglish.

Thames Valley Section

Held at: Reading Technical College.
Date: 10 November. Time: 7.15 p.m.
Lecture: Electronic Design.
By: H. V. Beck.

Merseyside Section

Held at: Walker Art Gallery, Liverpool.
Date: 17 November. Time: 6.30 p.m.
Lecture: Disc Recording and Production—Some Modern Developments.
By: D. G. Jacquess.

Scottish Section

Held at: University of Strathclyde, Glasgow
Date: 8 November. Time: 6 p.m.
Lecture: Telemetry—Present and Future Trends. (Joint meeting with the I.E.E.)
By: R. E. Young.
Held at: Carlton Hotel, North Bridge, Edinburgh 1.
Date: 9 November. Time: 6 p.m.
Lecture: Telemetry—Present and Future Trends. (Joint meeting with the I.E.E.)
By: R. E. Young.
Held at: University of Strathclyde, Glasgow.
Date: 23 November. Time: 6 p.m.
Lecture: The Use of Analogue Computers in Medicine.
(Joint I.E.E./I.E.R.E. Medical Electronics Group.)
By: W. J. Perkins.

SOCIETY OF ELECTRONIC AND RADIO TECHNICIANS

East and West Ridings Region

Held at: Hotel Metropole, King Street, Leeds 1.
Date: 8 November. Time: 7.15 p.m.
Lecture: Problems of Machine Tool Selection.
By: R. C. Brewer.

Midlands Region

Held at: Wolverhampton and Staffs College of Technology.
Date: 3 November. Time: 7.15 p.m.
Lecture: Production Control by Computer.
By: F. Pritchard.

South Eastern Region

Held at: King and Queen Hotel, 1 Marlborough Place, Brighton, Sussex.
Date: 10 November. Time: 7.30 p.m.
Lecture: Value Analysis.
By: A. Green.

Southern Region

Held at: College of Technology, Headington, Oxford.
Date: 9 November. Time: 7.30 p.m.
Lecture: The Applications of Computers in Engineering and Manufacturing Control at IBM Plants in Europe and the U.S.A.
By: D. H. Ralston.

Wales Region

Held at: Park Hotel, Park Place, Cardiff
Date: 9 November. Time: 7 p.m.
Discussion: Dynamic Gauging, Quality Control and Inspection.
(Members and invited guests only.)

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.
Date: 26 November.
Paper on: Television Audience Measurement.
By: D. J. Wheeler.

PUBLICATIONS RECEIVED

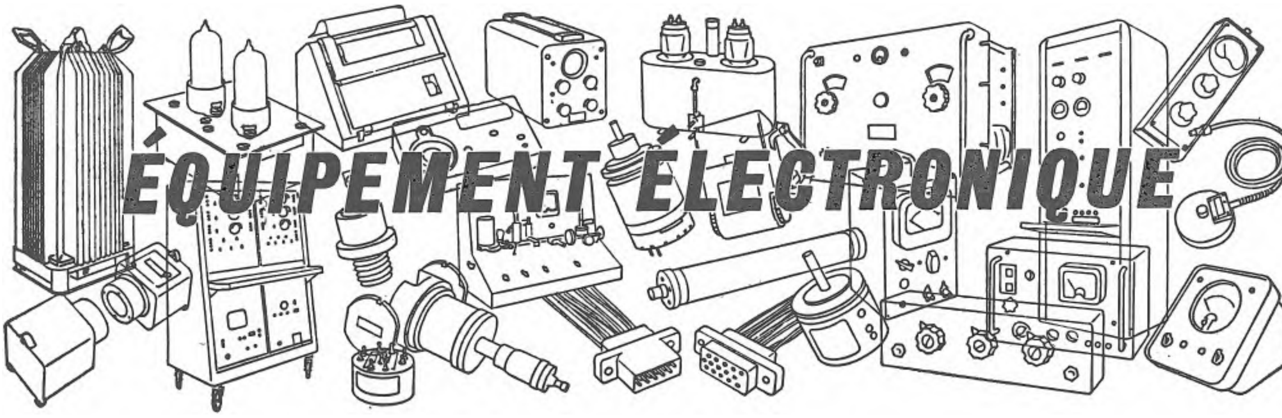
SEMICONDUCTOR INTEGRATED SOLID CIRCUITS is a 22-page booklet produced by Texas Instruments Ltd which is intended to facilitate the choice of the most suitable circuit for any given application. It provides concise details of the range of integrated circuits now available. Copies of this booklet are available from Texas Instruments Ltd, Manton Lane, Bedford.

ULTRASONICS REFERENCE HANDBOOK is a new technical service to industry inaugurated by Kelvin Electronics Company. This service takes the form of a bound loose-leaf handbook which gives a full technical appraisal of the fundamentals, technicalities and the scope within the field of ultrasonic non-destructive testing.

To date, the handbook has seven completed sections dealing with the basic principles of ultrasonics, ultrasonic shadow testing: direct echo testing: multiple echo testing: opacity testing: a guide to the correct probe selection and uses of a calibrated test block. The scope and contents of the handbook will be continuously reviewed and further parts or revised sections issued automatically to Registered Holders.

An initial registration fee of 5 guineas is made for this service, which covers the handbook; all amendments and additions, post free, which will be issued in the future.

Applications for registration should be made to: Kelvin Electronics Company, Publicity Department, Ref. CHO/RG, Kelvin House, Wembley Park Drive, Wembley, Middlesex.



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 762 à 767

SYSTÈMES À SOLÉNOÏDES SUPER-CONDUCTEURS

Distributeurs: Cryosystems Ltd, 40 Broadway, London, S.W.1

(Illustration à la page 762)

Une nouvelle gamme de systèmes super-conducteurs à solénoïdes pour produire des champs magnétiques élevés, fabriquée par la International Research & Development Co. Ltd, est offerte par la société Cryosystems Ltd.

Le principal avantage de ces systèmes est que, tout en retenant toutes les caractéristiques essentielles des appareils plus compliqués et plus coûteux, ils sont d'un fonctionnement simple et d'un prix réduit.

Le système de base, qui se compose d'un assemblage de solénoïdes, d'un syphon et d'accessoires et qui se trouvent centrés autour d'une solenoïde de 40 kG dont l'ouverture de travail est de 1,5 cm, peut être utilisé dans une variété étendue d'expériences de recherches et sur les corps solides.

Des systèmes à solénoïdes de conception spéciale peuvent être fournis également pour répondre aux besoins particuliers du client.

Notre gravure montre le bloc d'alimentation, la solénoïde standard et le cryostat portés par un support en bois dur. On voit également la solénoïde montée dans la position voulue.

EE 87 751 *pour plus amples renseignements*

CONTRÔLEUR DE RÉPONSE

Associated Electronic Engineering Ltd, Dalston Gardens, Stanmore, Middlesex

(Illustration à la page 762)

Cet instrument permet d'effectuer des variations infinies du niveau de signaux de fréquence acoustique par degrés d'octave de 40 Hz à 240 Hz.

Les neuf commandes séparées permettent chacune le réglage à variation continue de +13 dB à −13 dB.

Ce contrôleur qui a été conçu à l'intention des studios de radiodiffusion trouve également de nombreuses applications dans les studios d'enregistrement. Lorsqu'il est incorporé à un système de matériel de sonorisation de fabrique il permet de régler la réponse du système complet en vue d'une intelligibilité maxima.

Le modèle courant A1646 a des impédances d'entrée et de sortie de 600 Ω au niveau de 0 dBm et un affaiblissement d'insertion nul. Il est muni d'un commutateur d'arrêt-marche et d'une commande de gain. Le niveau de surcharge est de + 18 dBm. La réponse est de ±2 dBm environ entre 20 Hz et 20 kHz; le bruit est de −60 dB.

Le modèle standard est logé dans un coffret métallique avec commandes d'octaves éclairées dont les réglages sont indiqués sous forme graphique.

Il existe d'autres modèles pour montage sur bâti avec dispositifs d'entrée et de sortie à alternance.

EE 87 752 *pour plus amples renseignements*

OSCILLOSCOPE À DOUBLE TRACE

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

L'oscilloscope type 453 a été principalement conçu à l'intention des calculatrices constituées de corps solides et à action rapide. Il combine la performance et les caractéristiques d'un oscilloscope de laboratoire dans un coffret ne mesurant que 17,14 cm × 27,30 cm × 48,26 cm et pesant 12,69 kg. Il comprend un nouveau tube cathodique Tektronix de 10,16 cm étudié pour fournir les qualités d'enregistrement et de clarté exigées pour l'emploi dans des conditions lumineuses ambiantes élevées.

La sensibilité de la double trace est de 20 mV/div à 50 MHz et s'étend jusqu'à 5 mV/div à 40 MHz. Les canaux peuvent être mis en cascade pour obtenir une sensibilité de 1 mV/cm à 25 MHz sur trace unique. Le retard de signal permet la vision du bord principal de la forme d'onde de déclenchement.

Les quatre modes de fonctionnement comprennent le fonctionnement individuel de chaque canal, la commutation électronique à alternance ou par hachure entre les canaux ou l'addition algébrique des deux canaux. Dans le mode de fonctionnement à addition algébrique, le canal numéro 2 peut être inversé pour le fonctionnement différentiel.

Le dispositif de déviation horizontale comprend le retard de balayage étalonné pour des mesures pratiques telles que celles portant sur les intervalles entre impulsions et l'importance du vacillement sur les signaux de calculatrices ou n'importe quel train d'impulsions, différences de temps et angles de phase.

Le balayage étalonné s'étend de 5 sec/div à 0,1 μsec/div, avec un dispositif de grossissement portant le balayage le plus rapide à 10 nsec/div.

Les facilités de déclenchement à passe-bande totale comprennent le fonctionnement automatique, le balayage unique et le canal 1 seulement. Pour faciliter le fonctionnement, un voyant lumineux indique lorsque le balayage est déclenché.

Une image X-Y peut être obtenue lorsqu'on utilise le canal 1 pour l'entrée. Cet affichage fournit une sensibilité de 5 mV/div à 5 MHz sur l'axe horizontal.

Un calibreur, d'une précision de 1 % en amplitude et en fréquence, fournit des ondes carrées de 1 V et de 0,1 V, 1 kHz, pour faciliter l'étalonnage. Un circuit de sondage de courant de 5 mA est également fourni.

EE 87 753 *pour plus amples renseignements*

LECTEUR PHOTOÉLECTRIQUE À BANDE PERFORÉE

Great Northern Telegraph Works, 5 St. Helen's Place, London, E.C.3

(Illustration à la page 762)

Ce lecteur photoélectrique à bande perforée, modèle 4101, est fabriqué par la société Great Northern Telegraph Works du Danemark.

Il a été conçu pour servir d'élément de lecture pour calculatrice numérique électronique et pour d'autres appareils analogues exigeant un fonctionnement à grande vitesse. On peut faire varier la vitesse de fonctionnement continu de l'instrument de 250 à 1 000 caractères/sec et la vitesse de fonctionnement graduelle prévue est de 500 caractères/sec sur l'entrée spéciale pas-à-pas et de 1 000 caractères/sec avec la minuterie de processus (la durée depuis le début de l'ouverture de chaîne au nouvel ordre de marche) est de 60 μsec au minimum. La lecture est à réglage instantané pour les bandes à 5, 6, 7, ou 8 voies. La vitesse de la bande est contrôlée et peut être lue sur une échelle intérieure. La combinaison des caractères sur la station de lecture est indiquée sur le panneau de lecture.

Le lecteur a été conçu de façon à faciliter l'introduction de la bande car le verrou de la bande, lorsqu'il est ouvert, permet de mettre la bande en position avec précision sur la fenêtre de la station de lecture. A mesure que l'on baisse le verrou de la bande, trois doigts agrippent la bande et la maintiennent dans la position correcte jusqu'à ce que le frein entre en action lorsque le verrou de la bande est entièrement fermé. Le filetage de la bande à travers les cabestans, les rouleaux de guidage et la position de lecture ne sont pas nécessaires puisque le verrou de la bande place cette dernière automatiquement dans la position voulue pour quelle puisse se dérouler correctement. Un guide spécial de lumière assure l'éclairage uniforme de toutes les voies dans la bande. L'intensité lumineuse est maintenue à un niveau constant quelles que soient les fluctuations de la tension d'alimentation ou le vieillissement de la lampe qui est utilisée à 30 ou 40% en-dessous de sa puissance nominale afin de lui assurer une longue durée de vie, à l'aide d'un amplificateur asservi.

Il s'ensuit qu'un instrument fonctionnant à une telle vitesse de lecture est capable d'être arrêté et mis en marche très rapidement. Dans le modèle 4101, ces opérations s'effectuent au moyen de puissants électro-aimants. La bande peut être arrêtée en 0,8 msec et accélérée à une vitesse de lecture à partir de la position d'arrêt en 2 ou 3 msec environ.

Les amplificateurs de lecture à quatre étages fournissent des impulsions de sortie carrée avec un temps de montée et de chute inférieur à 2.0 µsec et à des longueurs qui dépendent de la vitesse de la bande. La sensibilité des amplificateurs peut être réglée intérieurement. L'appareil est également muni de trois amplificateurs d'entraînement pour les électro-aimants et d'un bloc d'alimentation stabilisé équipé de redresseurs au silicium. Les circuits sont à température compensée pour permettre le fonctionnement dans une gamme de température ambiante de 60°C. Un ventilateur à part assure un bon refroidissement de tous les composants. La spécification d'entrée et de sortie électronique rend le lecteur à bande compatible avec la plupart des appareils électroniques transistorisés.

Les circuits d'alarme donnent des renseignements à l'équipement électronique connexe en ce qui concerne le verrou de la bande, et la commande manuelle du déroulement de la bande.

Le bloc d'alimentation permet en outre le contrôle de la plupart des tensions.

L'instrument est prévu pour fonctionner sur tension alternative de 220 V, 50 Hz, ou 115 V, 60 Hz, et sa consommation électrique est d'environ 150 W. Il peut être fourni avec contrôle de parité des caractères moyennant un supplément de prix. Ce dispositif peut être mis en oeuvre en branchant deux plaquettes de circuit imprimé dans les connecteurs bobinés.

EE 87 754 *pour plus amples renseignements*

COMMUTATEURS DE PRESSION

Londex Ltd, 207 Anerley Road, London, S.E.20

(Illustration à la page 763)

La société Londex Ltd, qui est une filiale du groupe Elliott-Automation, fournit une nouvelle gamme de commutateurs de pression comportant à la fois un différentiel de réenclenchement à contact réglable et une gamme de fonctionnement réglable.

Il existe quatre modèles du nouveau type de commutateur PS/U qui couvrent des pressions de fonctionnement de 4,5 kg à 113,3 kg/6,45 cm². Les différentiels de réenclenchement réglable vont de 0,45 à 1,81 et de 1,36 à 4,5 kg/6,45 cm² suivant la gamme de pression du commutateur employé.

Ces nouveaux commutateurs de pression sont basés sur un soufflet en bronze phosphoreux sans soudure et à formation hydraulique qui fait oeuvre d'élément sensible à la pression. La dilatation ou la contraction du soufflet actionnent un commutateur à permutation unipolaire et à action rapide dont les caractéristiques sont comme suit: 5 A à 440 V c.a., 10 A à 250 V c.a., 0,2 A à 220 V c.c. non inductifs.

Les caractéristiques de fonctionnement de la gamme PS/U sont illustrées par l'exemple d'un commutateur à gamme réglable entre 13,60 et 45,35 kg/6,45 cm² ayant été réglé pour actionner ces contacts lorsque la pression tombe à 20,41 kg/6,45 cm². Vu que le différentiel de réenclenchement de ce dispositif peut être varié entre 0,45 et 2,72 kg/6,45 cm², les contacts dans la position minima se réenclencheront lorsque la pression s'élèvera à environ 20,86 kg/6,45 cm². Si la commande est calée au maximum, la pression de réenclenchement sera d'environ 23,12 kg/6,45 cm².

EE 87 755 *pour plus amples renseignements*

COMMUTATEURS NUMÉRIQUES

Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(Illustration à la page 763)

Une nouvelle gamme de commutateurs numériques a été réalisée par la société Digitizer Techniques Ltd. Les commutateurs "Digiswitch" ont été conçus pour les applications exigeant la lecture en ligne des chiffres. La construction modulaire alliée à une plaquette de commutation à circuit imprimé a permis de créer un élément d'une épaisseur de 1,27 cm grâce auquel on peut effectuer des densités de tassement élevées. Le montage des contacts est normalement de dix ou douze directions unipolaires mais des contacts à deux directions et cinq pôles où cinq directions bipolaires peuvent être prévus sur demande. Les contacts ont été conçus pour la commutation du courant alternatif ou du courant continu jusqu'à 50 V à 100 mA, mais l'indice de courant nominal continu peut s'élever à 1 A. Ces puissances nominales ont été obtenues par l'emploi d'un contact standard au graphite d'argent en liaison avec un circuit imprimé doré, dont la couche sous-jacente est normalement en papier impregné de résine synthétique mais qui peut être également en verre de fibre sur demande spéciale.

Le commutateur se caractérise en particulier par l'éclairage par l'arrière des caractères. Ce dispositif peut être ajouté comme supplément facultatif. Des ampoules de différentes tensions peuvent être fournies, dont notament des ampoules de 6, 12 et 24 V.

EE 87 756 *pour plus amples renseignements*

ENREGISTREUR À BANDE À SIX VOIES

Hewlett Packard, Dallas Road, Bedford

(Illustration à la page 763)

Six voies de données peuvent être reportées en même temps grâce au nouvel enregistreur à six voies, modèle 7190C. Prévu pour les mesures industrielles ou scientifiques, le modèle 7190C utilise six cartouches à encre actionnées par solénoïdes et montées dans un seul assemblage qui élimine pratiquement les parties mobiles caractéristiques des instruments conventionnels. Il est entièrement transistorisé et il a été conçu pour l'emploi sans le moindre ennui et sans surveillance aucune.

Une série de facilités de contrôle et offerte aux utilisateurs qui doivent contrôler des opérations et produire des enregistrements visuels. Ces options comprennent un potentiomètre de contrôle qui se déplace en même temps que la tête enregistreuse ainsi que trois commutateurs de limites pour la commande arrêt-marche. Ces options peuvent être fournies avec l'enregistreur ou ajoutées à ce dernier par la suite.

Il existe une variété de modules d'entrée couvrant des gammes de température de 100° C à 1600° C, des gammes de tension de 5 mV à 250 V et des gammes de courant de 250 mA à 5 A. D'autres modules d'entrée peuvent être fournis sur commande spéciale.

EE 87 757 *pour plus amples renseignements*

AMPLIFICATEUR ET INDICATEUR DE MILLIVOLTS

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustration à la page 764)

La société Kent Precision Electronics a réalisé un nouveau millivoltmètre dans sa gamme d'instruments de contrôle industriel constitués de corps solides.

L'instrument se compose d'un amplificateur à transistors KPE à couplage direct et à faible dérive qui alimente un ampèremètre de type standard. L'ensemble de l'instrument est incorporé dans un coffret standard mesurant 8,89 cm, pour montage sur panneau.

L'amplificateur convient pour une gamme d'entrée effective minima de 0 à

10 mV et l'entrée peut être également adaptée pour relier des éléments de résistance sous forme de pont afin d'assurer une indication précise. L'instrument contient normalement une tension de référence précise. L'amplificateur a une sortie de courant de 0 à 1 mA avec une impédance de sortie très élevée et il peut être utilisé pour alimenter des indicateurs à distance externes, ainsi que des enregistreurs et des contrôleurs, en plus de l'ampèremètre indicateur intérieur.

Cet instrument trouve de nombreuses applications pour l'indication de température, l'enregistrement ou le contrôle par l'emploi de thermocouples (pour lesquels la compensation automatique par jonction à froid peut être prévue) ou de thermomètres à résistance. Il peut être utilisé avec la gamme KPE d'instruments de contrôle constitués de corps solides, car il fournit un signal de courant permettant d'actionner un instrument de contrôle étalonné standard KPE constitué de corps solides, en plus de sa fonction normale d'indicateur.

EE 87 758 *pour plus amples renseignements*

RÉCEPTEUR DE COMMUNICATIONS

Redifon Ltd, Communications Division, London, S.W.18

(Illustration à la page 764)

La société Redifon Ltd a mis au point un récepteur de communications transistorisé d'une stabilité totale et d'une sensibilité élevée. Ce récepteur a été conçu pour la téléphonie universelle, la télégraphie et la réception fac-similée dans la gamme de 13 kHz à 28 MHz.

L'appareil Redifon R.408, dont le modèle a été approuvé par l'administration des téléphones, convient en effet pour les services de point à point, pour les stations mobiles ou fixes et pour l'usage maritime. Des normes élevées de fiabilité sont obtenues grâce à l'emploi étendu de transistors planaires au silicium.

Le R.408 a une largeur de bande à variations continues et il peut être utilisé pour la réception sur ondes porteuses, sur modulation d'amplitude, sur bande latérale unique ou n'importe quelle bande latérale indépendante (sélection par commutation de la bande latérale inférieure ou supérieure). Le contrôle automatique de gain peut être effectué à des temps d'attaque ou de décroissance convenant pour tous les services. Un nouveau type de circuit conçu par Redifon assure le temps d'attaque rapide et le temps de décroissance prolongée exigés pour la réception sur bande porteuse et sur bande latérale unique, sans le désavantage de la "paralysie" par éclatements de bruit ou interférences.

Des facilités sont prévues pour la réception sur bande latérale unique (porteuse pilote ou entièrement supprimée) ainsi que pour n'importe quelle bande de transmission par bande latérale indépendante. La largeur de bande est à variation continue de 800 Hz à 8 kHz pour la réception à modulation d'amplitude et sur onde porteuse, et de 800 Hz à 4 kHz pour la réception sur bande latérale unique. Lorsque l'appareil est utilisé sur onde porteuse, une largeur de bande étroite de 160 Hz peut être commutée au moyen d'un filtre à cristal.

Le dispositif de contrôle a été conçu sur la base des résultats d'un grand nombre d'opérations pratiques. Une échelle d'accord à film de 228 cm rend superflu l'emploi d'une échelle de concentration séparée, cependant que l'étalonnage par cristal s'effectue à intervalles de 100 kHz.

Des sorties de fréquences acoustiques, de fréquences moyennes en ligne et de contrôle de gain automatique sont prévues; l'appareil est, en outre, entièrement protégé contre les court-circuits et contre les signaux de tout transmetteur connexe, même lorsque le R.408 est débranché.

L'appareil R.408 standard fonctionne sur alimentations de 100 à 125 V ou 100 à 250 V, 50 à 60 Hz c.a., et les modèles spéciaux fonctionnent sur alimentations en courant continu de 24 V, 110 V ou 220 V.

EE 87 759 *pour plus amples renseignements*

POTENTIOMÈTRE DE PRÉCISION

H. Tinsley & Co. Ltd, Werndee Hall, South Norwood, London, S.E.25

(Illustration à la page 764)

La société H. Tinsley & Co. Ltd vient d'ajouter à sa gamme de potentiomètres un instrument de précision en mesure de surmonter les limites du potentiomètres à courant continu de type classique. Le potentiomètre "Stabaumatic", type 5545, s'inspire d'un nouveau principe de construction basé sur le contrôleur de courant automatique Tinsley, type 5388. Ce principe rend l'instrument indépendant de la stabilité du courant de batterie et il peut donc être utilisé de façon immédiate. La stabilité est de trois parties dans un million. Pour les travaux de haute précision, une batterie indépendante à cellule standard et à température contrôlée peut être utilisée.

Un circuit à potentiel de résistance constant ne comportant aucun contact de commutation est utilisé. Il y a en outre, sept cadrans de mesure sous forme de conductances avec degrés binaires. Le choix de n'importe lequel des trois circuits de mesure s'effectue au moyen d'un commutateur à cinq directions. La quatrième direction sur le commutateur choisit la cellule standard intérieure et la cinquième position convertit l'instrument en un pont de mesure à résistance à six gammes. Un réglage de cellule standard à trois cadrans, relié à travers le circuit de potentiel, permet de normaliser le potentiomètre à une partie dans 10^6 de la force électromagnétique d'une cellule standard interne ou externe.

L'étalon habituel à résistance de réaction dans le circuit de contrôle de courant automatique est remplacé par les éléments de conductance à commutation du potentiomètre, le courant total passant à travers ces éléments étant série avec le circuit de potentiel. potentiel fourni est proportionnel courant, ce dernier pouvant être va en actionnant les conductances au mo des cadrans de mesure. La fuite de t sion entre les conductances commut dans le circuit de mesure est mainte constante et égale à la force électrom nétique d'une cellule standard au mo du contrôleur de courant automatiq Lorsque les conductances sont mises oeuvre, le courant dans le circuit potentiel à résistance constante est va de manière à ce que les indicati des cadrans de conductance soient éga à la tension parcourant le circuit. Ce tension est équilibrée en fonction d'u tension inconnue au moyen du détect galvanométrique de la manière ha tuelle. Les conductances débranchées circuit de mesure sont reliées de fa à ce quelles fassent toujours partie parcours de courant total.

Toutes les conductances sont co mutées en parallèle par les cadrans p produire des valeurs de décade bien le nombre des bobines ne soit que quatre bobines par cadran. Ce mont facilite le réglage et élimine les erre de résistance étant donné que les c tacts se trouvent dans des circuits résistance relativement élevée et qu sont en parallèle. Nulle erreur thermi ne peut être causée par la commutat dans le circuit de potentiel, ce der ayant une résistance constante de 10 sur la gamme supérieure et de 10Ω la gamme inférieure. On a donc potentiomètre exempt d'erreurs ther ques pour les travaux de thermocoupl à basse tension de haute précision.

La normalisation de l'instrum s'effectue par le réglage des cadrans mesure à la valeur connue d'une cell standard se trouvant à l'intérieur l'instrument et par l'ajustement du positif de réglage de la cellule stand afin d'assurer l'équilibre correct. Une mise au point, la normalisation maintenue par l'une des cellules inter équilibrant le contrôleur. Le potent mètre comprend en outre deux cellu de commande. L'une de ces cellules comme cellule intermédiaire dur l'équilibrage préliminaire au cours duq les courants déséquilibrés peuvent r verser temporairement l'équilibre de cellule. La cellule intermédiaire et deuxième cellule de commande ou contrôle sont choisies par le commu teur de contrôle de sensibilité, de so que la cellule de travail n'est mise oeuvre que lorsqu'un équilibre rigoure a été obtenu. Une troisième cellule st dard est utilisée pour la normalisati La compensation de températu s'effectue automatiquement pour variations de température ambiante 19° C à 24° C. A condition que la te pérature ambiante demeure dans ce gamme, la normalisation ne doit ê vérifiée en fonction de la cellule inte qu'une ou deux fois par jour.

Deux gammes de mesure sont prévu à savoir celle de 0 à 2,11111 V par pl μV et celle de 0 à 211,111 mV par s de 0,1 μV. A l'intérieur de ces mes, l'instrument a une précision onnée de ±0,001% de l'indication à μV près et une limite d'erreur à long ıe de ±0,003% de l'indication à μV près. La gamme de l'instrument :ant que pont s'étend de 0,01 Ω à 1Ω et la limite d'erreur est de 0,005% 0,0005 Ω dans la gamme de 0,01 Ω kΩ, puis de 0,01% jusqu'à 211 kΩ, e 0,1% jusqu'à 2,1 MΩ et de 1% à MΩ. Pour cette dernière application, :ontrôleur automatique de courant . pas nécessaire, mais il constitue partie essentielle du circuit lorsque trument est utilisé comme potentio- e.

: 87 760 *pour plus amples renseignements*

TRANSDUCTEURS DE DÉPLACEMENT

tersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustration à la page 765)

:s transducteurs de déplacement ntiométriques type DR68 construits la société Intersonde Ltd sont prévus des longueurs de course de 1,27 2,54 cm et 5 cm et ils sont basés e principe d'un élément de résistance née de précision. Dans ces trans- ɜurs, le déplacement rectilinéaire de sensible actionne un frotteur en .l noble sur l'élément de résistance ɜst normalement excité par une ten- constante. La tension du frotteur directement proportionnelle au ɪcement appliqué et, en cas de ɪcement total, la tension du frotteur ɪubstantiellement égale à la tension :itation. Avec une excitation appro- ɪ, la sortie du transducteur est suffi- :' pour actionner de nombreux types licateurs et d'enregistreurs sans ne amplification ne soit nécessaire.

:s transducteurs constituent un moyen coûteux et sûr de percevoir le ɪcement et ils offrent une précision rieure à 0,5% de la gamme totale et résolution allant jusqu'à 0,001 :e. Les limites de température de tionnement s'étendent de —20° C à .0° C et la durée de vie des trans- eurs dépasse 1 000 000 d'opérations.

: 87 761 *pour plus amples renseignements*

COMPARATEUR DE BANDE

Г. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustration à la page 765)

: comparateur de bande "Aptec," èle AT 9102, est un des plus petits, plus rapides et des plus indépendants e genre d'instrument. Il constitue une :on modifiée du lecteur de bande ɔis G.N.T. à deux têtes et il peut parer des bandes perforées de 5, 6, 7 : niveaux dans n'importe quel régime :ode. Conçu pour contrôler la pré- ɪn de copies de bandes produites à ir de bandes originales, il a une se maxima de 30 codes par seconde. : comparateur Aptec actionne les les originales et les copies devant comparées au moyen d'un axe à moteur commun et à travers deux têtes de lecture commandées séparément. Lorsque le codage perçu par les deux têtes s'accorde exactement, les dispositifs d'entraînement vers les deux porte-bande sont excités simultanément et les deux bandes "passent" à la position de code suivante. Lorsque des codes dissemblables sont perçus, l'entraînement est supprimé et les deux bandes s'arrêtent. Après vérification des différences, les bandes peuvent être alignées à nouveau sur le code identique suivant au moyen de boutons d'avance normale ou à un seul coup. Lorsque les codes s'accordent de nouveau, le comparateur continu les contrôle automatiquement. Les deux têtes de lecture sont munies de dispositifs d'interverrouillage pour la tenue des bandes qui remettent la machine à l'arrêt en cas de défaut de bande. Des interverrouillages analogues entrent en action si l'une ou l'autre des têtes de lecture est laissée ouverte. Les interverrouillages détectent également tout endommagement aux ouvertures de galet des bandes.

Le comparateur Aptec se distingue par la facilité de la charge, qui est accomplie en posant simplement les bandes sur les galets dans la porte ouverte de chaque tête. Les portes sont ensuite fermées et la machine est prête à fonctionner. Les boutons d'avance normale ou à un coup permettent l'alignement très rapide des bandes avant la comparaison.

La machine peut être utilisée avec de grosses collures, y compris les jointures au chlorure de polyvinyle. Elle constitue un appareil indépendant mesurant 46,9 cm × 20,3 cm × 17,7 cm.

EE 87 762 *pour plus amples renseignements*

ALIMENTATIONS STABILISÉES

Kingshill Electronic Products Ltd, Torrens Works, Torrens Street, London, E.C.1

(Illustration à la page 765)

Les blocs d'alimentation stabilisée de la série G constituent des éléments à modulateur de haute performance prévus pour une gamme de tension de 0 à 30 V et un courant maximum de 10A.

Le coût de ces blocs a pu être réduit au minimum grâce à la rationalisation électrique et mécanique de la construction. Ils comportent des composants de haute qualité tels que des transformateurs à noyau C et des résistances à haute stabilité.

La construction mécanique judicieusement étudiée de ces blocs permet de les monter de n'importe quel côté et en particulier de les monter dans une position autre que la verticale habituelle. On peut donc les monter sur la face inférieure. Ces caractéristiques sont d'un avantage très net car elles offrent une variété de méthodes de montage qui facilitent le remplacement de blocs de marque différente.

Deux dimensions de panneau de base sont prévues, à savoir: 13,01 cm × 10,16 cm et 13,01 cm × 13,97 cm. La longueur des éléments varie de 6,98 cm à 38,1 cm assurant ainsi la plus grande souplesse d'empilage derrière un panneau standard de 48,26 cm. Les dimensions modulaires s'étendent de 13,01 cm × 10,16 cm × 6,98 cm pour une puissance nominale de 0 à 30 V ½ A ou 0 à 15 V, 1 A, à 13,01 cm × 13,97 cm × 38,1 cm pour une puissance nominale de 0 à 15 V, 10 A.

Chaque bloc comprend un limiteur de courant électronique à réenclenchement automatique qui peut être préréglé permettant un déplacement d'environ 5 V. Le dispositif de perception à distance est de type normal et le rail positif est relié à un maillon pour le contrôle de courant ou l'alarme.

La régulation est supérieure à 0,1 %, le rapport de stabilisation est de 1000:1, le bourdonnement est de 1 mV, la tolérance de courant secteur est de ± 10 % et la température ambiante maxima de 45°C.

EE 87 763 *pour plus amples renseignements*

RELAIS DE PUISSANCE

Magnetic Devices Ltd, Newmarket, Suffolk

(Illustration à la page 766)

Un relais bipolaire et à deux directions de construction robuste et conçu pour allier la fiabilité maxima à l'économie maxima a été créé par la société Magnetic Devices Ltd. Convenant pour les connexions de soudage rapide, ces terminaisons ont été groupées ensemble de manière pratique dans un coffrage moulé. En outre, ses barrières à arc et un espacement généreux entre les parties sous tension en font un instrument des plus utiles pour les applications les plus difficiles.

Les relais, prévus pour courant alternatif ou courant continu, sont protégés par une housse transparente Makrolon, et les matériaux de moulage spécial utilisés dans l'appareil garantissent ses propriétés de résistance à l'arc. Le montage s'effectue au moyen d'un seul goujon relié à un ergot de localisation.

Le relais de la série 140 EP (c.a.) est d'une valeur nominale de 10 A et sa puissance nominale de bobine est de 5,5 VA, sa tension de régime maxima est de 140 V et sa résistance de bobine maxima de 10 kΩ. Le relais de la série 145 EP (c.c.) est d'une puissance nominale de 10 A; sa puissance de bobine nominale est de 1,75 W, sa tension de régime maxima est de 150 V et sa résistance de bobine maxima de 14 kΩ. (Une résistance série peut être fournie séparément pour les relais de la série 145 EP pour des tensions supérieures.)

Les vitesses moyennes de fonctionnement et de déclenchement des relais série 145 EP sont de 25 Msec. Ils peuvent être utilisés dans des températures ambiantes atteignant 35° C et ils peuvent être adaptés pour des températures plus élevées sur demande.

Le poids maximum du relais est de 187 grammes et ils peuvent être munis de contacts en argent fin ou en elkonite. Enfin, il existe des modèles avec bobines imprégnées au vide pour les conditions tropicales ou humides.

EE 87 764 *pour plus amples renseignements*

CONTRÔLEURS DE TEMPÉRATURE

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

(Illustration à la page 766)

Une nouvelle série de contrôleurs de température de précision à thermocouples, portant la référence de série 18-90, a été mise sur le marché par la société Ether Ltd. Il existe actuellement quatre type de contrôleurs de cette série.

Ces instruments comprennent trois sections principales: un point de réglage de précision, un amplificateur de déviation et un contrôleur. Ils peuvent être mis en action sur alimentation normale de secteur dans des températures ambiantes de 0 à 50°C.

Sur les types 18-92 et 18-98, un commutateur à bouton poussoir à sept positions et un potentiomètre de précision à 10 tours assurent le calage du point de réglage. Chacune des positions de commutation fournit des plots de 100° et le potentiomètre divise chacun des plots de 100° en mille divisions. On obtient ainsi une précision de réglage de 0,1°. Le signal d'entrée est filtré puis comparé au signal de sortie de la section de point de réglage. Le signal d'erreur qui en résulte est injecté à l'amplificateur de déviation dont la sortie est communiquée à un instrument de mesure de la déviation. Sur le type 18-92, la sortie indiquée sur l'enregistreur de déviation est transmise à la section de contrôle qui produit un signal de 0 à 5 mA c.c. Sur le type 18-98, la sortie peut produire des impulsions d'amorçage pour l'entraînement de thyristors.

Les types 18-99 et 18-92 sont identiques aux types 18-92 et 18-98 respectivement, à l'exception du fait qu'ils ont été conçus comme contrôleurs de thermocouples différentiels. A la place du point de réglage numérique, ils comportent des échelles étalonnées avec 50 divisions égales. L'étalonnage normal est de $-0{,}5$ mV à $+0{,}5$ mV.

EE 87 765 *pour plus amples renseignements*

MOTEUR SYNCHRONE À FAIBLE VITESSE

Vactric Control Equipment Ltd, Garth Road, Morden, Surrey

(Illustration à la page 766)

Une nouvelle version du moteur synchrone à faible vitesse "Vacsyn" vient d'être mise au point aux usines de Morden de la société Vactric Control Equipment Ltd. Ce nouvel appareil a été conçu pour les applications exigeant une vitesse réduite et constante, telles que les machines à copier de bureau, la régulation des tubes de porte ou la commande de guidage des machines-outils.

Pour faciliter l'installation dans les espaces restreints, le diamètre hors-tout du moteur a été réduit à moins de 10 cm mais la puissance de sortie originale de 6,6 W a été conservée. Sous sa forme standard, le moteur est bobiné pour 115 V, 50 Hz, donnant un couple minimum de 4,25 kg/2,5 cm. Il existe des versions dont la vitesse est de 60 ou 71, 4 tours/min. Des bobinages et des vitesses différentes peuvent être prévus sur commande spéciale.

La vitesse synchrone réduite du moteur est obtenue par l'emploi d'un rotor à plusieurs pôles à aimantation permanente dont le stator comporte un bobinage biphasé équilibré. Ce moteur fonctionne sur alimentation uniphasée, le déphasage du second bobinage étant assuré par une résistance série externe et un condensateur.

Le moteur n'exige aucun appareillage coûteux et il est, en outre, réversible et mis en marche automatiquement. Il peut être utilisé comme moteur élévateur par l'application d'une séquence d'impulsion négative et positive à chaque bobinage. Tout changement de polarité fait avancer l'axe de 1,8° ou de 2,14° selon le type du moteur, de sorte qu'il faut 200 ou 168 plots pour compléter un tour.

EE 87 766 *pour plus amples renseignements*

MICRORUPTEUR

The Plessey Co. Ltd, Electromechanical Division, Havant, Hampshire

(Illustration à la page 766)

Le nouveau microrupteur subminiature que vient de créer la Division Electromécanique du Groupe des Composants Plessey allie une haute sensibilité à une longue durée de vie mécanique.

Le nouveau microrupteur type 19 comprend, en effet, un mécanisme à lame de commutation qui ne comporte que deux parties mobiles et dont le mouvement différentiel de fonctionnement se réduit à 0,0005pouce au maximum pour l'action ultra-rapide et sensible.

Sa durée de vie mécanique dépasse 10 millions de cycles tandis que la durée de vie électrique à pleine charge nominale de 5 A à 250 V c.a. est de 100 000 Hz. La durée de vie électrique nominale a demicharge est de 250 000 Hz.

Le microrupteur type 19 consiste essentiellement en un élément unipolaire à deux directions mais il peut également être fourni sous forme d'élément à une direction normalement ouverte et normalement fermée. Les bornes sont soudées et du type à barrillet. Le matériau de contact standard est l'argent. On peut également obtenir sur demande des contacts en argent doré ou en matérieux spéciaux.

La force de fonctionnement normal du microrupteur type 19 est de 117,2 à 141,5 grammes et il peut être muni d'une gamme complète de commandes. Le coffret prévu pour les applications courants mesure 0,25 pouce × 0,78 pouce × 0,36 pouce; il consiste en un moulage de résine phénolique et un matériau en diallyl peut être fourni pour l'utilisation jusqu'à 200°C.

EE 87 767 *pour plus amples renseignements*

COMMUTATEUR ROTATIF

Honeywell Controls Ltd, Brentford, Middlesex

(Illustration à la page 766)

La société Honeywell Controls Ltd a réalisé un nouveau commutateur rotatif d'une grande souplesse d'emploi qui p remplacer jusqu'à huit tumblers.

Cet élément compact, qui porte référence 1RM, n'occupe que 1, pouce × 1,45 pouce d'espace de panne Il est de construction robuste et b sur les commutateurs Honeywell V3 grande capacité. Ces commutateurs p vent être utilisés dans n'importe qu séquence—jusqu'à un maximum 4000 variations- et le même élém peut loger des commutateurs utilisal à différents niveaux de courant, contacts dorés étant utilisés pour circuits à faible énergie et des cont d'argent pour les circuits de 15 A. commutateurs sont actionnés par arbre à cames, construit pour fou une séquence de précision pendant très longue durée de temps, le fo tionnement se faisant à la main ou moteur.

L'un des avantages du nouvel élém consiste dans l'action rapide des co mutateurs V3 qui éliminent les err de bruit blanc qui se produisent ha tuellement avec les commutateurs galette rotatifs. L'élément est scellé d le panneau à l'aide d'un joint à 1 adhésive et d'un anneau "O".

EE 87 768 *pour plus amples renseignem*

AMPLIFICATEURS SÉLECTIFS

Avon Communication and Electronics Lt 14-16 High Street, Christchurch, Hampshi

(Illustration à la page 767)

La nouvelle gamme d'amplificate sélectifs accordables, série 400, d' grande souplesse d'emploi et à bas p a été prévue pour une variété éten d'applications dans les domaines ind triel et de télécommunication.

Ces amplificateurs sont à accord c tinu dans une gamme de fréquence 270 Hz à 370 Hz sur quatre bandes. peut faire varier la sélectivité entre et 20% et un bon rejet des fréquen adjacentes à la passe-bande est ass par l'emploi de deux circuits simulta ment accordés donnant une caracté tique d'atténuation approchant 120 octave.

Les éléments actifs sont transistor et actionnés par batterie interne donn une durée de vie d'environ trois m Les amplificateurs sont de construct robuste et compacte et ils sont mu d'un instrument de mesure de 3,81 pour l'accord des signaux.

Lorsque l'amplificateur est utilisé d les réseaux de télécommunication, rendement est amélioré par l'emp d'un étage de sortie non-linéaire "porte", car en imposant un seuil niveau de signal minimum à la sor on obtient une nette amélioration rapport signal/bruit.

Cet instrument peut améliorer le r port signal/bruit d'une liaison de t communication d'au moins 20 dB.

Dans les applications de laborato telles que la détection nulle de pont, étages de sortie linéaire peuvent fixés et les amplificateurs sélectifs peuv améliorer considérablement la précis points nuls. Des modèles peuvent : fournis avec circuits limiteurs incorés, pour obvier au choc auditif ;qu'ils sont utilisés pour les applicais de pont et la détection par écouteur.

E 87 769 *pour plus amples renseignements*

UNISÉLECTEURS À LAME VIBRANTE

Alma Components Ltd, Park Road, Diss, Norfolk

(*Illustration à la page 767*)

:es nouveaux unisélecteurs à lame t des éléments de faibles dimensions, ;, souples et considérablement moins teux que d'autres dispositifs de comtation électronique. Un unisélecteur nplet à trois bancs et dix directions mesure que 8,89 cm × 8,89 cm × ! cm. Il ne nécessite qu'une alimenon de 24 V et avance ensuite d'une ition lorsque deux fils métalliques t réunis; aucun circuit supplémentaire :t nécessaire.

.e fonctionnement inversé peut être ctué et on peut ajouter un dispositif itoguidage à n'importe quel chiffre sélectionné. On peut également inporer à l'unisélecteur un dispositif mettant de le relier en décade à itres éléments par le "report" à zéro ı élément à l'autre. N'importe quel ıbre de directions, et ce jusqu'à 100, t être prévu avec un maximum de bancs. Des contacts de permutation des contacts normalement fermés vent être fournis à la place des cons normalement ouverts de type ıdard.

'emploi d'un système à lame offre de nombreux avantages: faible consommation électrique (½ W), grande vitesse de fonctionnement contact entièrement scellé n'exigeant aucun entretien ou réglage et d'une longue durée de vie, contacts indépendants à la place du circuit frotteur ordinaire et une résistance à faible contact.

EE 87 770 *pour plus amples renseignements*

THERMOMÈTRE ÉLECTRONIQUE

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(*Illustration à la page 767*)

Le thermomètre électronique type 166C constitue le dernier-né de la gamme des indicateurs de température à thermocouples.

Le type 166C prévu pour l'emploi avec des thermocouples cromel/alumel de n'importe quelle résistance de conducteur, peut lire la température directement et sans cartes, de −120° C à +1100° C avec une résolution d'un degré à toutes les températures.

La jonction froide automatique, l'amplificateur à relais modulateur constitué de corps solides et les batteries au mercure de longue durée assurent une mesure rapide et précise avec un minimum de commandes.

La tension de sortie des thermocouples est compensée en plots de 100° C par un diviseur de potentiel de précision étalonné en fonction d'une cellule de référence interne. L'enregistreur couvre les intervalles de 100°. Une gamme de recherche de 100 à 1000° C est également prévue.

La sortie de l'enregistreur standard est de 200 μA.

EE 87 771 *pour plus amples renseignements*

CONTRÔLEUR DE SEMI-CONDUCTEURS

Distributeurs: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(*Illustration à la page 767*)

Le "Diotestor", fabriqué par la société Sercel S.A.R.L. de France, à été spécialement conçu pour le contrôle des semi-conducteurs sur circuit imprimé à l'extrémité de la chaîne de montage. Il détecte les transistors et diodes défectueux sans qu'il soit nécessaire de déssouder ses composants.

Le "Diotestor" vérifie les conditions de conduction et de coupure des transistors et diodes pnp/npn. Le processus de contrôle est simple et n'exige aucune connaissance spéciale.

Deux modèles sont prévus, dont l'un comporte une batterie sèche standard de 3 V pesant 150 grammes et l'autre des accumulateurs Voltabloc (rechargeables par branchement direct sur le courant secteur) et pesant 311 grammes.

Le "Diotestor" est d'un emploi facile et pratique en raison de son poids réduit et de sa forme maniable. Il est logé dans un coffret incassable en Rislan et doté d'une alimentation propre.

Deux accessoires sont prévus, un adapteur à sept espacements différents pour le contrôle des diodes et un câble à rallonge avec douille et pince crocodile pour le contrôle des semi-conducteurs non-montés.

EE 87 772 *pour plus amples renseignements*

Résumés des Principaux Articles

charge de bruit blanc de systèmes de communications multivoies par W. Oliver

Résumé de l'article aux pages 714 à 717

L'auteur examine une méthode d'essai utilisée mondialement pour le contrôle des systèmes de communications multivoies. Il montre que le bruit gaussien a les mêmes propriétés statistiques que le signal multivoies et indique des méthodes pour la conversion des résultats obtenus par les mesures du rapport de puissance de bruit en fonction du rapport bruit/signal.

bruit de grenaille et la mesure de la charge électronique par E. Mathieson

Résumé de l'article aux pages 718 à 721

L'auteur donne des détails expérimentaux de la mesure du bruit de grenaille dans le courant d'une diode thermoélectronique.

Cette mesure a permis d'obtenir une valeur pour la charge électronique de $1.62.10^{-19}$ coulombs, l'incertitude expérimentale de ce résultat étant d'environ 5%. Le rôle de ce type de recherche en tant qu'expérience de laboratoire pour étudiants est examiné par l'auteur qui donne également un aperçu de la théorie qui s'y rapporte.

non-linéarité des résistances fixes par P. L. Kirby

Résumé de l'article aux pages 722 à 726

L'emploi d'un coefficient de tension pour définir l'importance de la non-linéarité des résistance fixes comporte de gros désavantages et l'auteur suggère une autre méthode qui comporte la mesure de la distorsion harmonique. Dans un cas simple on a pu produire une sortie de troisième harmonique proportionnel au cube d'une tension fondamentale appliquée et obtenir, sur cette base, un nouvel élément appellé "index de troisième harmonique". L'article indique l'origine probable de la non-linéarité, ainsi que la raison de sa corrélation avec le bruit de courant. Les valeurs caractéristiques d'indexe de troisième harmonique sont indiquées pour les résitance à film dans une gamme étendue de valeurs. Les applications futures de l'index comprennent l'évaluation de la performance d'une résistance de base pour les spécifications d'essai. Il peut, en outre, être employé comme méthode de blindage pour assurer un degré très élevé de fiabilité.

Un inverseur de 1 KW fonctionnant sur fréquence secteur par W. D. Gilmour

Résumé de l'article aux pages 727 à 731

Cet article décrit un inverseur fournissant jusqu'a 1 KW à des fréquences connues avec précision entre 45 et 65 Hz et dont la tension de sortie est parfaitement stable. La sortie est sinusoïdale à des charges allant jusqu'a 300W et devient graduellement plus carrée à mesure qu'on augmente les charges. Un système de commande à faible tension est utilisé.

Les circuits de commutation à transistors au silicium à couplage direct par J. D. Martin

Résumé de l'article aux pages 732 à 735

L'utilisation des circuits de commutation à transistors au silicium à couplage direct est bien connue mais il existe très peu de données d'étude à son sujet. Le problème a été étudié afin de réaliser des circuits sûrs plutôt que des circuits à gain maximum. Cet article analyse donc la tension de saturation du collecteur à transistor et indique la manière de calculer les paramètres d'étude. Les expériences effectuées avec les circuits conçus selon cette méthode ont confirmé l'utilité du procédé.

Un générateur de précision triphasé à 400Hz par D. J. Miles

Résumé de l'article aux pages 736 à 739

On a réalisé une alimentation triphasée, stable en amplitude et en fréquence, servant d'alimentation de puissance pour les gyroscopes à inertie. Cette alimentation assure une stabilité d'amplitude de ±0,01% par rapport à la charge normale et aux variations de courant secteur.

L'étude des amplificateurs à l'aide d'une calculatrice numérique par J. R. Abrahams

Résumé de l'article aux pages 740 à 745

Cet article indique la manière dont on peut utiliser une petite calculatrice numérique pour l'étude et le calcul de circuits électroniques. La réalisation d'un amplificateur accordé à deux étages et à transistors est utilisée comme exemple numérique, avec une certaine simplification des paramètres. L'article analyse également la possibilité d'effectuer des calculs comparatifs de prix et de réaliser ainsi la construction la plus économique possible.

Une méthode itérative pour résoudre les problèmes des mailles multiples par N. M. Morris

Résumé de l'article aux pages 745 à 746

Cet article traite d'une méthode itérative pour résoudre les problèmes des mailles multiples. Cette méthode est basée sur la série de Taylor et elle peut être appliquée à n'importe quel réseau linéaire de tension continue ou de tension alternative ainsi qu'à quelques réseaux non linéaires. Elle constitue une variante utile à la solution des réseaux à mailles multiples en l'absence d'un simulateur.

Un amplificateur de rapport transistorisé utilisant le principe du calcul de temps écoulé par A. A. Cremin et M. C. Sexton

Résumé de l'article aux pages 747 à 748

L'auteur décrit un circuit analogique qui permet d'obtenir le rapport de deux signaux d'entrée variant dans le temps. Cette méthode comporte l'emploi de circuits de temps écoulé (base de temps) pour produire des fréquences d'échantillonnage considérablement supérieures aux fréquences des signaux d'entrée. Il en résulte un signal de sortie essentiellement continu et proportionnel au rapport d'entrée.

Résistance de sortie négative dans un bloc d'alimentation régulée par J. Diamond

Résumé de l'article aux pages 749 à 751

Une résistance de sortie négative est parfois souhaitable dans une alimentation de puissance régulée. Elle peut être utilisée pour annuler la résistance de conducteurs ou d'un composant quelconque en série avec la charge, dans les cas où la solution habituelle, c'est à dire l'échantillonnage à distance de la tension de charge, ne peut être appliquée.

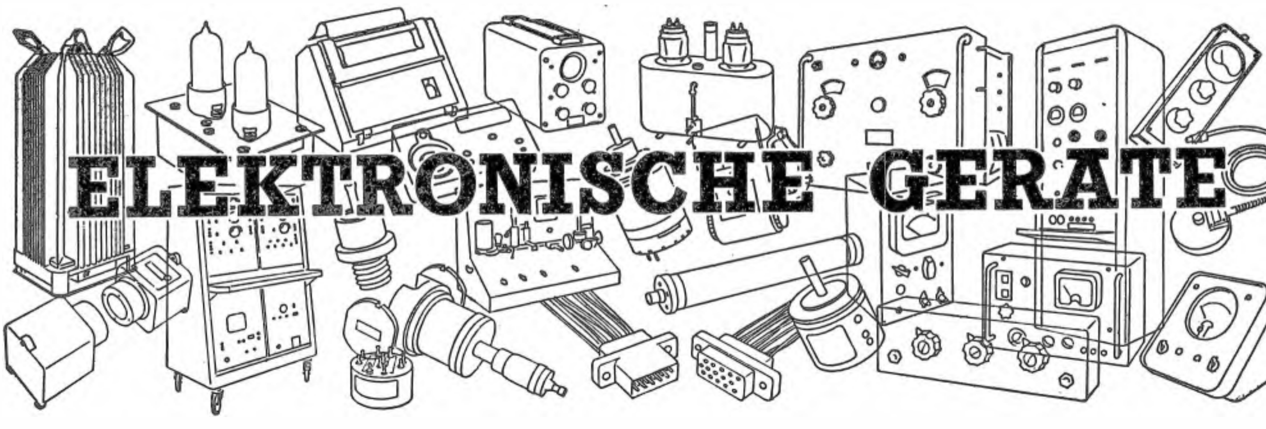
La conception d'une pareille alimentation est examinée dans cet article, qui montre également qu'on peut obtenir une résistance de sortie négative constante, indépendamment de la tension de sortie. On peut, en outre, contrôler le coefficient de température de la résistance de sortie.

La détermination de la matrice d'impédance de réseaux linéaires compliqués par W. A. Crabbe

Résumé de l'article aux pages 752 à 753

L'auteur analyse la méthode Kron qui permet d'obtenir la matrice d'impédance d'un réseau linéaire. Cette méthode est ensuite appliquée à un simple amplificateur tripolaire puis à un amplificateur à deux étages à réaction en pont. Les résultats de certaines configurations de réseaux relativement compliquées sont simplifiés par l'emploi de la matrice d'impédance. La méthode Kron pour obtenir cette matrice constitue une technique précise produisant moins d'erreurs.

ELEKTRONISCHE GERÄTE



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben

Übersetzung der Seiten 762 bis 767

Supraleitende Solenoidsysteme
Vertrieb: Cryosystems Ltd, 40 Broadway, London, S.W.1
(Abbildung Seite 762)

Die von International Research & Development Co. Ltd herausgebrachten neuen supraleitenden Solenoidsysteme für die Erzeugung starker Magnetfelder können nunmehr von Cryosystems Ltd bezogen werden.

Einer der Hauptvorteile der neuen Systeme ist, dass sie einfach und preiswert sind, trotzdem alle wesentlichen Eigenschaften der aufwendigeren und teureren Ausrüstungen beibehalten wurden.

Das Grundsystem besteht aus dem Solenoid-Zusammenbau, dem Kryostat und Ständer, Stromversorgung, Heber und Zubehör und hat als Mittelpunkt ein 40-kG-Solenoid mit einer Arbeitsöffnung von 38,1 mm. Es hat Anwendungsmöglichkeiten in einer grossen Anzahl von Festkörperexperimenten und anderen Forschungsarbeiten.

Solenoidsysteme können spezifischen Kundenwünschen entsprechend in Sonderkonstruktion ausgeführt werden.

Die Abbildung zeigt die Stromversorgung, das Standardsolenoid und den Kryostat auf einem Hartholzständer mit dem Solenoid in Position.

EE 87 751 *für weitere Einzelheiten*

Frequenzgang-Kontrollgerät
Associated Electronic Engineers Ltd, Dalston Gardens, Stanmore, Middlesex
(Abbildung Seite 762)

Das Gerät gestattet kontinuierliche Regelung des Tonfrequenz-Signalpegels in Oktavenstufen von 40 ... 10 240 Hz. Jeder der neun getrennten Regler erlaubt stufenlose Änderung des Pegels von +13 dB ... −13 dB.

Das für Rundfunkstudios entwickelte Gerät hat auch viele Anwendungsmöglichkeiten in Aufnahmestudios. Einbau in die Betriebslautsprecheranlage ermöglicht Einregeln des Gesamtfrequenzganges des Systems auf optimale Sprachdeutlichkeit.

Das Standardmodell A1646 hat Eingangs- und Ausgangsimpedanzen von 600 Ω bei einem Pegel von 0 dBm und keine Durchgangsdämpfung. Ein Ein-Ausschalter und Verstärkungsregler sind vorhanden. Der Überlastungspegel ist +18 dBm. Der Frequenzgang ist zwischen 20 Hz und 20 kHz innerhalb ±2 dBm, Rauschen −60 dB.

Das Standardmodell ist in einem Metallgehäuse untergebracht, die Oktavenregler sind beleuchtet, und ihre Einstellung wird in grafischer Form dargestellt.

Andere Modelle sind für Gestelleinbau sowie mit alternativen Ein- und Ausgangsanordnungen lieferbar.

EE 87 752 *für weitere Einzelheiten*

Zweistrahl- Oszillograf
Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

Der Typ 453 wurde hauptsächlich für den Service von schnellen Festkörper-Elektronenrechnern an Ort und Stelle entwickelt und kombiniert Leistung und Eigenschaften eines Laboroszillografen in einem Gerät mit Abmessungen von nur 171 × 273 × 483 mm und etwa 12,7 kg Gewicht. Er ist mit der neuen Tektronix-4″-Elektronenstrahlröhre bestückt, die die bei hellen Umgebungsbedingungen erforderliche Helligkeit und hohe Schreibgeschwindigkeit hat.

Die Zweistrahlempfindlichkeit ist von 20 mV/Teilung bei 50 MHz bis 5 mV/Teilung bei 40 MHz, und durch Kaskadenschaltung der Kanäle kann man eine Empfindlichkeit von 1 mV/cm bei 25 MHz im Einstrahlbetrieb erzielen. Signalverzögerung gestattet Beobachtung der Vorderflanke der Triggerungswellenform.

Es gibt vier Betriebsmethoden, und zwar jeder Kanal einzeln, beide Kanäle abwechselnd, oder zerhacktes elektronisches Schalten zwischen den Kanälen, oder beide Kanäle algebraisch zusammenaddiert. In der letzteren Betriebsweise kann Kanal 2 für Differentialbetrieb umgekehrt werden.

Die Horizontalablenkung hat zur bequemen Ausführung von Messungen wie z.B. Impuls-zu-Impuls-Abstand und dem Zittern an Rechnersignalen oder beliebigen Impulsreihen, Zeitdifferenzen und Phasenwinkeln, eine geeichte Ablenkverzögerung.

Geeichte Ablenkgeschwindigkeiten erstrecken sich von 5 s/Teilung bis zu 0,1 μs/Teilung mit einer 10fachen Beschleunigung der schnellsten Ablenkung auf 10 ns/Teilung.

Komplette Triggerungseinrichtungen über den vollen Durchlassbereich umfassen u.a. automatischen Betrieb, Einmalablenkung und "nur Kanal 1". Zur weiteren Erleichterung der Bedienung leuchtet eine Lampe auf, wenn die Ablenkung getriggert wird.

Man kann Kanal 1 für den X-Eingang verwenden und erhält dann eine X-Y-Anzeige. Die Empfindlichkeit der horizontalen Achse ist dann 5 mV/Teilung bei 5 MHz.

Ein Eicher, dessen Amplituden- und Frequenzgenauigkeit innerhalb einer 1-Prozent-Toleranz liegt, gibt zur bequemen Eichung 1-kHz-Rechteckimpulse bei 1 V und 0,1 V ab. Eine Schleife für einen 5-mA-Strommesskopf ist auch vorhanden.

EE 87 753 *für weitere Einzelheiten*

Fotoelektrischer Streifenleser
Great Northern Telegraph Works, 5 St. Helen's Place, London, E.C.3
(Abbildung Seite 762)

Der in beiden Richtungen arbeitende fotoelektrische Lochstreifenleser wird von den Great Northern Telegraph Works in Dänemark hergestellt.

Das Gerät trägt die Modellnummer 1401 und wird hauptsächlich für die Eingabe in elektronische Digitalrechner und ähnliche Anlagen, die hohe Lesegeschwindigkeit erfordern, eingesetzt. Bei kontinuierlichem Betrieb kann die Geschwindigkeit zwischen 250 und 1 000 Zeichen/Sek. geregelt werden; sie ist bei schrittweisem Ablesen über eine schrittweise Spezialeingabeeinheit mit einer Lesegeschwindigkeit von 500 Zeichen/Sek. und einer Prozesszeit (das ist die Zeit zwischen dem Anfang des Streifenloches bis zum Startbefehl) von mindestens 60 μs bis zu 1 000 Zeichen/Sek. lieferbar. Das Ablesen kann augenblicklich auf Streifen mit 5, 6, 7 oder 8 Kanälen umgestellt werden. Die Streifengeschwindigkeit wird überwacht und auf einer internen Skala angezeigt. Die über der Lesestation stehende Zeichenkombination ist im Anzeigefeld sichtbar.

Das Einlegen des Streifens wird durch die Konstruktion des Lesers erleichtert, da das genaue Positionieren des Streifens über dem Lichtfenster der Lesestation bei offener Streifenführung möglich ist. Bei Senken der Streifenführung halten drei Finger den Streifen in der korrekten Lage, bis bei völligem Schliessen der Streifenführung die Bremsen diese Aufgabe übernehmen. Da diese Streifenführung den Lochstreifen automatisch für korrekten Vorschub positioniert, braucht der Streifen nicht zwischen Treib- und Führungsrollen und der Lesestation eingelegt zu werden. Eine Spezialeinrichtung gewährleistet gleichmässige Beleuchtung aller Kanäle des Streifens. Die Lichtstärke wird auch bei Netzschwankungen oder Alterung der Lampe, die für längere Lebensdauer 30...40 Prozent unter Nennlast läuft, durch Einsatz eines Servoverstärkers konstant gehalten.

Starten und Stoppen eines mit so hoher Geschwindigkeit laufenden Gerätes muss sehr schnell erfolgen. Im Modell 1401 wird das durch sehr leistungsfähige Elektromagnete erreicht. Der Lochstreifen kann in 0,8 ms gestoppt und von der Stopposition in etwa 2...3 ms auf Lesegeschwindigkeit gebracht werden.

Der vierstufige Leseverstärker gibt Rechteckimpulse mit Anstiegs- und Abfallzeiten von 2,0 μs und von der Streifengeschwindigkeit abhängender Dauer ab. Die Empfindlichkeit des Verstärkers lässt sich intern regeln. Ausserdem gibt es noch drei Treiberverstärker für die Elektromagnete und eine mit Siliziumzellen bestückte Konstantstromversorgung. Die Schaltungen sind temperaturkompensiert und gestatten Betrieb in einem Umgebungstemperaturbereich von 60° C. Ein getrenntes Gebläse gewährleistet gute Kühlung aller Bauelemente. Die technischen Daten des elektronischen Eingangs und Ausgangs geben dem Streifenleser Kompatibilität mit den meisten transistorisierten elektronischen Ausrüstungen.

Schutzschaltungen geben angeschlossenen elektronischen Geräten Information in bezug auf offene Streifenführung, Streifenenden und manuelle Kontrolle des Streifenvorschubs.

Die meisten Spannungen der Stromversorgung können überwacht werden.

Das Gerät kann entweder für 220 V~ 50 Hz oder 115 V~ 60 Hz geliefert werden und hat eine Energieaufnahme von etwa 150 W. Es ist gegen Zuschlag mit Zeichenparitätskontrolle erhältlich, die durch Einstecken von zwei Druckschaltungsplatten in bereits verdrahtete Steckverbindungen nachzurüsten ist.

EE 87 754 *für weitere Einzelheiten*

Druckschalter

Londex Ltd, 207 Anerley Road, London, S.E.20

(Abbildung Seite 763)

Londex Ltd, ein zur Elliott-Automation-Gruppe gehörendes Unternehmen, hat neue Druckschalter herausgebracht, in denen sowohl das Rückstelldifferential für die Kontakte, als auch der Arbeitsbereich regelbar sind.

Von den neuen Schaltern des Types PS/U gibt es vier Modelle, die Betriebsdrücke von 0,7 ... 17,6 kp/cm² überstreichen. Die einstellbaren Rückstelldifferentiale hängen vom Druckbereich des betreffenden Schalters ab und liegen zwischen 0,07...0,3 und 0,2...0,7 kp/cm².

Dem neuen Druckschalter liegt eine hydraulisch geformte nahtlose Druckdose aus Phosphorbronze zugrunde, die als druckempfindliches Element wirkt. Ausdehnen oder Zusammenziehen der Druckmessdose betätigt einen einpoligen Schnapp-Umschalter, der für die folgenden typischen Belastungen bemessen werden kann: 5 A bei 440 V~, 10 A bei 250 V~, 0,2 A bei 220 V— induktionsfrei.

Die Betriebsdaten des PS/U-Fertigungsprogrammes werden beispielsweise durch einen Schalter illustriert, dessen Arbeitsbereich von 2,1 bis zu 7 kp/cm² überstreicht und für Betätigung der Kontakte bei Fallen des Druckes unter 3,16 kp/cm² eingestellt ist. Da sich das Rückstelldifferential für dieses Modell zwischen 0,07 und 0,42 kp/cm² einregeln lässt, werden die Kontakte bei Einstellung auf die niedrigste Position bei 3,23 kp/cm² zurückgestellt. Wenn die Steuerung auf den Höchstwert eingestellt ist, wird der Rückstelldruck bei etwa 3,51 kp/cm² liegen.

EE 87 755 *für weitere Einzelheiten*

Daumenradschalter

Digitizer Techniques Ltd, 26 Sheen Road, Richmond, Surrey

(Abbildung Seite 763)

Digitizer Techniques Ltd hat eine Serie von Daumenradschaltern herausgebracht. Der mit "Digiswitch" bezeichnete Schalter wurde besonders für Anwendungszwecke entwickelt, bei denen einzeilige Anzeige eingestellter Ziffern erforderlich ist. Modulare Konstruktion ergibt zusammen mit einer Schalterplatte in Drucktechnik einen nur 12,7 mm breiten Baustein, so dass hohe Packdichten erreicht werden können. Kontaktanordnungen sind normalerweise einpolig mit 10 oder 12 Positionen, können aber auch 5polig mit 2 Positionen oder 2polig mit 5 Positionen geliefert werden. Die Kontakte wurden für Gleich- oder Wechselstromschalten von bis zu 50 V bei 100 mA bemessen, die Nennbelastung kann jedoch bei Dauerbetrieb 1 A sein. Diese Belastbarkeit wurde mit einem Standard-Silber-Graphitkontakt auf einer hartvergoldeten Leiterplatte erzielt, deren Träger normalerweise aus Hartpapier, im Sonderauftrag jedoch aus Glasfaserkunststoff besteht.

Ein ungewöhnliches, gegen Zuschlag erhältliches Konstruktionsmerkmal ist die Beleuchtung der Schriftzeichen von hinten. Für die Lampen können verschiedene Spannungen vorgeschrieben werden, z.B. 6, 12 oder 24 V.

EE 87 756 *für weitere Einzelheiten*

Sechskanal-Streifenschreiber

Hewlett Packard, Dallas Road, Bedford

(Abbildung Seite 763)

Sechs Informationskanäle können mit dem neuen Sechskanal-Regelschreiber 7190 gleichzeitig aufgetragen werden. Das Modell 7190 ist für industrielle und wissenschaftliche Messungen mit sechs, durch Magnetspulen betätigten Tintenpatronen in einem Einzelzusammenbau ausgerüstet, der effektiv alle herkömmliche Konstruktionen kennzeichnenden beweglichen Teile beseitigt. Das volltransistorisierte Gerät ist für unbemannten, wartungsfreien Betrieb ausgelegt.

Benutzern, die ausser dem Protokoll Regelvorgänge benötigen, wird wahlweise eine Anzahl von Regeleinrichtungen angeboten, zu denen ein Regelpotentiometer, das sich mit dem Registrierkopf bewegt, sowie bis zu drei Grenzschalter für Ein-Ausschalten gehören. Diese wahlweisen Zusätze können mit dem Schreiber geliefert oder nachgerüstet werden.

Die verschiedensten Eingangsmoduln sind lieferbar und überstreichen Temperaturspannen von 100° C bis zu 1600° C, Spannungsbereiche von 5 mV bis zu 250 V und Strombereiche von 250 mA bis zu 5 A. Andere Eingangsmoduln sind gegen Sonderauftrag lieferbar.

EE 87 757 *für weitere Einzelheiten*

Millivoltverstärker und -anzeiger

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Abbildung Seite 764)

Kent Precision Electronics Ltd hat im Rahmen ihres Programmes für industrielle Halbleiter-Kontrollinstrumente KPE ein neues Millivolt-Anzeigegerät herausgebracht.

Das Instrument besteht aus dem galvanisch gekoppelten Transistorverstärker KPE mit niedriger Drift, der einen Standardstromanzeiger zugelassenen Types speist. Das gesamte Gerät ist in einem Standard-Instrumentgehäuse für Frontplattenmontage und 89 mm Durchmesser untergebracht.

Der Verstärker ist für einen effektiven Mindesteingangsbereich von 0...10 mV geeignet; der Eingang kann auch für Anschluss von Widerständen in Brückenschaltung zur Erzielung genauer Anzeigen adaptiert werden. Normalerweise enthält das Gerät eine genaue Bezugsspannung. Der Verstärker hat bei sehr hoher Ausgangsimpedanz einen Ausgangsstrom von 0...1 mA und ist zusätzlich zum internen Anzeigeinstrument für Speisung externer Fernanzeiger, Schreiber und Regler geeignet.

Das Gerät hat breite Anwendungsmöglichkeiten für Temperaturanzeige, -registrierung und -regelung entweder mit Thermoelementen (wenn automatische Kompensation der Kaltlötstellen-Temperatur vorgesehen werden kann) oder mit Widerstandsthermometern. Es kann mit anderen KPE-Festkörper-Kontrollinstrumenten eingesetzt werden und gibt dann ausser der normalen Anzeigefunktion ein Stromsignal ab, das ein geeichtes KPE - Standard - Festkörper - Kontrollinstrument betreiben kann.

EE 87 758 *für weitere Einzelheiten*

Nachrichtenempfänger

Redifon Ltd. Communications Division, London, S.W.18

(Abbildung Seite 764)

Redifon Ltd hat einen hochempfindlichen und konstanten volltransistorisierten Nachrichtenempfänger entwickelt, der für Allzweckfernsprech-, Telegrafie- und Bildübertragungsempfang im Frequenzbereich 13 kHz...28 MHz ausgelegt ist.

Die mit Redifon R.408 bezeichnete Ausrüstung ist von den britischen Postbehörden zugelassen und für Funkverkehr zwischen zwei festen Punkten, ortsfeste oder mobile Stationen und Marineverwendung geeignet. Durch umfangreiche Anwendung von Silizium-Planartransistoren wird eine hochgradige Zuverlässigkeit erzielt.

Der R.408 hat eine kontinuierlich regelbare Bandbreite und kann für den Empfang von CW, AM, ESB oder jedem beliebigen Seitenband bei 2 SB (umschaltbare Wahl des oberen oder unteren Seitenbandes) eingesetzt werden. Volle wählbare AVR ist mit für alle Dienste geeigneten Einregel- und Abklingzeiten vorhanden. Eine neue, von Redifon entwickelte Schaltungsart gibt die für CW und ESB erforderlichen schnellen Einregelzeiten und langsamen Abklingzeiten ohne den Nachteil der durch Stösse, Rauschen oder Störungen hervorgerufenen "Paralyse."

Einrichtungen für vollen ESB (Pilotträger oder voll unterdrückter Träger) und für jedes der Seitenbänder bei Doppeleinseitenbandsendungen sind vorhanden. Die Bandbreite ist zwischen 800 Hz und 8 kHz für AM und CW, zwischen 800 Hz und 4 kHz für ESB kontinuierlich regelbar. Bei CW-Betrieb lässt sich über ein Kristallfilter eine Schmalbandbreite von 160 Hz einschalten.

Die Regelorgane wurden nach breiter Untersuchung der Betriebspraxen ausgelegt. Eine 229 cm lange Film-Abstimmskala macht eine getrennte Aufzeichnungsskala überflüssig; Kristalleichung gibt es in 100-kHz-Abständen.

Tonfrequenz, Leitungs-ZF und AVR können entnommen werden, und der Empfänger ist gegen Stromkreisunterbrechung, Kurzschluss und Signale zugehöriger Sender selbst im abgeschalteten Zustand voll geschützt.

Der Standardempfänger R.408 kann mit 100...125 V~ oder 200...250 V~, 50...60 Hz betrieben werden; Sondermodelle sind für 24 V–, 110 V– oder 220 V– lieferbar.

EE 87 759 *für weitere Einzelheiten*

Präzisionskompensator

H. Tinsley & Co. Ltd. Werndee Hall, South Norwood, London, S.E.25

(Abbildung Seite 764)

H. Tinsley & Co Ltd hat neuerdings ihr Kompensatorprogramm durch ein Präzisionsinstrument ergänzt, das die Begrenzungen herkömmlicher Gleichspannungskompensatoren überkommt. Das als "Stabaumatic" Typ 5545 angebotene Instrument enthält ein neues Gerätekonzept, das sich auf den Automatischen Stromregler Tinsley-Typ 5388 stützt und das Instrument von der Batteriekonstanz unabhängig und damit sofort betriebsbereit macht. Die Stabilität ist 3×10^{-6}. Für Aufgaben, die höchste Präzision erfordern, kann man eine unabhängige, temperaturgeregelte Normalelementbatterie anwenden.

Eine Konstantwiderstand-Spannungsschaltung ohne Schaltkontakte wird benutzt, und es gibt sieben Messskalen in Leitwertform und binären Stufen. Jede beliebige der drei Messschaltungen kann mittels eines 5stelligen Drehschalters eingestellt werden. Die vierte Schalterstellung stellt die Verbindung zum internen Normalelement her, und die fünfte verwandelt das Instrument in eine Widerstandsmessbrücke mit sechs Messbereichen. Ein Normalelementeinsteller mit drei Skalen liegt parallel zur Spannungsschaltung und gestattet Standardisieren des Kompensators auf 1×10^{-6} der EMK des internen oder externen Normalelementes.

Der übliche Rückkopplungs-Normalwiderstand in der Schaltung der Stromregelautomatik wird durch die geschalteten Konduktanzen des Kompensators ersetzt, und der Gesamtstrom fliesst durch diese Konduktanzen in Serie mit der Spannungsschaltung. Die erzeugte Spannung ist dem Strom, der mit der Veränderung des Leitwertes mittels der Messskalen schwankt, proportional. Der Potentialabfall an diesen Konduktanzen, die in die Messkreise geschaltet werden, wird mittels der Stromregelautomatik konstant gehalten und entspricht der EMK eines Normalelementes.

Bei Bedienung der Konduktanzen wird der Strom in der Konstantwiderstand-Spannungsschaltung so geändert, dass Konduktanzskalen die an der Schaltung liegende Spannung anzeigen. Diese Spannung wird in üblicher Weise mittels eines Galvanometers gegen eine unbekannte Spannung abgeglichen. Die aus dem Messkreis geschalteten Konduktanzen bilden weiterhin einen Teil des Gesamtstromweges.

Alle Konduktanzen werden zur Erzeugung von Dekadenwerten durch die Skalen parallelgeschaltet, trotzdem es nur vier Spulen je Skala gibt; dadurch wird das Justieren vereinfacht und Widerstandsfehler vermieden, da die Kontakte in verhältnismässig hochohmigen Schaltungen und parallel liegen. In der Spannungsschaltung, die im oberen Bereich einen konstanten Widerstand von 100 Ω und im unteren Bereich einen von 10 Ω hat, werden keine thermischen Fehler durch Schalten hervorgerufen. Daraus ergibt sich ein von thermischen Störungen freier Kompensator für äusserst genaues Arbeiten mit niederspannigen Thermoelementen.

Standardisierung des Instrumentes wird durch Einstellen der Messskalen auf den bekannten Wert des eingebauten Normalelementes bewirkt; der Normalelementeinsteller wird dann so lange verändert, bis der richtige Abgleich hergestellt ist. Danach wird die Standardisierung aufrecht erhalten, da eins der Normalelemente, den Regler kompensiert. Es gibt zwei interne Kontrollelemente, von denen eins während eines Vorabgleiches, wenn Messdiagonalströme den Elementabgleich zeitweise stören können, als Pufferelement dient. Das Pufferelement und das zweite oder Arbeitskontrollelement werden durch einen Empfindlichkeitsreglerschalter gewählt, so dass das Arbeitselement nicht eingeschaltet wird, bis der Feinabgleich beginnt. Ein drittes Normalelement wird für das Standardisieren benutzt. Änderungen der Umgebungstemperatur zwischen 19° C und 24° C werden automatisch kompensiert. Bleibt die Umgebungstemperatur in diesen Grenzen, so ist Standardisierung gegen das interne Element nur ein- oder zweimal am Tag notwendig.

Es gibt zwei Messbereiche: von 0...2,11111 V in Stufen von 1 μV und von 0...211,111 mV in Stufen von 0,1 μV. Das Gerät hat über diese Bereiche eine geeichte Messgenauigkeit von ±0,001 Prozent der Ablesung +1 μV und eine Langzeitfehlergrenze von ±0,003 Prozent der Anzeige +1 μV. Als Brücke hat das Gerät einen Messumfang von 0,01 Ω ... 21 MΩ mit einer Messfehlergrenze von ±0,005 Prozent im Bereich 0,01 Ω... 21 kΩ, dann 0,01 Prozent bis zu 211 kΩ, 0,1 Prozent bis zu 2,1 MΩ und 1 Prozent bis zu 21 MΩ. Für diesen Anwendungszweck ist die Stromregelautomatik nicht erforderlich; für Einsatz des Gerätes als Kompensator bildet sie jedoch einen unerlässlichen Teil des Instrumentes.

EE 87 760 *für weitere Einzelheiten*

Weggeber

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Abbildung Seite 765)

Der von Intersonde Ltd hergestellte potentiometrische Weggeber Typ DR68 beruht auf einem Präzisions-Drahtwiderstandselement und ist mit 12,7 mm, 25,4 mm und 51 mm Hub lieferbar. In diesen Messwertgebern treibt eine geradlinige Verschiebung des Fühlschaftes einen

Edelmetallschleifer über ein Widerstandselement, das normalerweise durch eine konstante Spannung erregt wird. Die Schleiferspannung ist der Fühlerverschiebung direkt proportional, und bei voller Nennverschiebung entspricht die Schleiferspannung im wesentlichen der Erregerspannung. Bei geeigneter Erregung ist der Ausgang des Messwertgebers hoch genug, um—wo erforderlich—Anzeiger oder Registriergeräte ohne Verstärkung zu treiben.

Die Messwertgeber sind ein zuverlässiges und preisgünstiges Mittel zur Erfassung von Verschiebungen und bieten eine Unsicherheit von besser als 0,5 Prozent des Messendwertes und eine Auflösung von 0,0254 mm. Der Betriebstemperaturbereich erstreckt sich von −20° C bis zu +120° C, und die Lebensdauererwartung überschreitet 1 Million Betätigungen.

EE 87 761 *für weitere Einzelheiten*

Lochstreifenvergleicher

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 765)

Der "Aptec"—Lochstreifenvergleicher AT9102 ist eins der kleinsten und schnellsten in sich geschlossenen lieferbaren Geräte dieser Art. Er enthält eine modifizierte Ausführung des dänischen Doppelkopf-Streifenlesers der G.N.T. und kann 5-, 6-, 7- oder 8-Kanal-Lochstreifen in jedem Kode vergleichen.

Das zur Überprüfung der von Originalen kopierten Lochstreifen auf Fehlerfreiheit entwickelte Gerät arbeitet mit einer Höchstgeschwindigkeit von 30 Kodezeichen je Sekunde.

Der "Aptec"-Vergleicher treibt Original und zu vergleichende Kopie über eine gemeinsame Motorwelle durch zwei getrennt gesteuerte Lesestationen. Wenn der von beiden Lesern abgetastete Kode genau übereinstimmt, wird der Antrieb der beiden Streifentransporte gleichzeitig erregt, und beide Streifen werden in die nächste Kodeposition vorgeschoben. Stimmen die gelesenen Kodes nicht überein, so werden die Antriebe gesperrt und beide Streifen stoppen. Nach Überprüfung der Unstimmigkeit können die Lochstreifen durch Drücken der "Vorwärts"- oder "Einmal"-Tasten auf die nächsten übereinstimmenden Kodezeichen ausgerichtet werden. Wenn die Zeichen übereinstimmen, läuft der Vergleicher automatisch weiter. Beide Lesestationen haben Verriegelungseinrichtungen, die das Gerät stoppen, wenn der Streifen klemmt oder ausläuft. Ähnliche Verriegelungseinrichtungen kommen in Betrieb, wenn eine Lesestation offen bleibt. Die Kontrolle "Streifen aus" kann auch beschädigte Perforationslöcher erkennen.

Besonders einfach ist beim "Aptec"-Vergleicher das Einlegen der Lochstreifen, die nur im offenen Fenster jeder Lesestation über die Zahntrommel gelegt werden. Das Fenster wird dann geschlossen, und das Gerät ist betriebsfertig. Die "Einmal"- und "Vorwärts"-Tasten erleichtern das Ausrichten der Streifen vor Beginn des Vergleiches.

Der Betrieb der Ausrüstung, die bei Abmessungen von 470 × 203 × 178 mm ein in sich geschlossenes Gerät darstellt, wird durch dicke Klebestellen einschliesslich solcher mit PVC-Streifen nicht beeinflusst.

EE 87 762 *für weitere Einzelheiten*

Konstantstromversorgungen

Kingshill Electronic Products Ltd, Torrens Works, Torrens Street, London, E.C.1

(Abbildung Seite 765)

Die Konstantstromversorgungen der Baureihe G sind modulare Hochleistungseinheiten für den Spannungsbereich 0...30 V und bis zu 10 A Höchststrom.

Durch Rationalisieren der elektrischen sowie mechanischen Ausführung wurden die Kosten gesenkt. Bauelemente bester Qualität finden durchweg Verwendung, z.B. Transformatoren mit Schnittbandkern und hochkonstante Widerstände.

Die saubere Konstruktion gestattet Montage der Einheiten auf jeder beliebigen Fläche; ausserdem brauchen sie nicht aufrecht installiert zu werden, man kann sie sogar hochkant einbauen. Dadurch bieten sich bei Auswechseln anderer Fabrikate die verschiedensten Montagemöglichkeiten, was ein besonderer Vorteil ist.

Die Geräte kommen in zwei Frontplattengrössen: 130 × 102 mm und 130 × 140 mm; die Länge der Einheiten liegt zwischen 70 und 381 mm, so dass für Anordnung hinter der Frontplatte eines 19"-Gestelleinschubs die grösstmöglichste Anpassungsfähigkeit besteht. Die Abmessungen der Moduln liegen für eine Belastbarkeit von 0...30 V, 0,5 A oder 0...15 V, 1 A zwischen 130 × 102 × 70 mm, für 0...15 V, 10 A bis zu 130 × 140 × 381 mm.

Jedes Gerät hat einen selbstrückstellenden elektronischen Strombegrenzer, dessen Arbeitspunkt über einen grossen Bereich vorgewählt werden kann, und einen voreingestellten Ausgangsspannungsregler, der ungefähr 5 V Pendeln zulässt. Fernabtastung gehört zur Standardausführung, und die positive Schiene wird zu einer Verbindung für Stromüberwachung oder ein Warnsystem herausgeführt.

Diese Regelung ist besser als 0,1 Prozent, das Regelverhältnis 1000:1, die Restwelligkeit 1 mV, Netztoleranz ±10 Prozent und maximale Umgebungstemperatur 45°C.

EE 87 763 *für weitere Einzelheiten*

Hochbelastbares Relais

Magnetic Devices Ltd, Newmarket, Suffolk

(Abbildung Seite 766)

Ein robustes zweipoliges Umschaltrelais, in dem höchste Zuverlässigkeit mit grösster Wirtschaftlichkeit verbunden ist, wurde von Magnetic Devices herausgebracht. Die Anschlüsse sind in bequemer Weise in einem gepressten Gehäuse gruppiert und für Löten oder "Schnellverbindungsbuchsen" geeignet; durch die ausreichenden Lichtbogenschranken und reichlichen Abstände zwischen stromführenden Teilen ist das Relais für Einsatz unter schwierigen Bedingungen geeignet.

Das für Gleich- und Wechselstrom lieferbare Relais hat eine durchsichtige Makrolon-Haube, und die durchweg verwendeten Spezialpressmassen geben guten Schutz gegen Kriechströme. Montage erfolgt mittels eines Einzelgewindestiftes 3BA in Verbindung mit einem Führungsansatz.

Baureihe 140EP (Wechselstrom) ist für 10 A bemessen, hat eine Spulennennleistung von 5,5VA, Höchstbetriebsspannung von 440 V und Höchstspulenwiderstand von 10 kΩ. Baureihe 145EP (Gleichstrom) ist für 10 A bemessen, hat eine Spulennennleistung von 1,75 W, Höchstbetriebsspannung von 150 V und Höchstspulenwiderstand von 14 kΩ. (Für Baureihe 145EP ist getrennt ein Serienwiderstand für höhere Spannungen lieferbar).

Die durchschnittlichen Ansprech- und Abfallzeiten der Baureihe 145EP sind 25 ms. Das Relais kann in Umgebungstemperaturen bis zu 35°C arbeiten und auf Wunsch für höhere Temperaturen modifiziert werden.

Das maximale Gewicht ist 227 g; Feinsilber- oder Elkonit-Kontakte sind lieferbar. Ausserdem gibt es vakuumimprägnierte Spulen für feuchte oder tropische Betriebsbedingungen.

EE 87 764 *für weitere Einzelheiten*

Temperaturwächter

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

(Abbildung Seite 766)

In der neu vorgestellten Baureihe 18-90 der Ether Ltd gibt es zur Zeit vier Präzisions - Thermoelement - Temperaturwächter.

Diese Temperaturinstrumente bestehen aus drei Teilen, und zwar einem Präzisions-Sollwertgeber, einem Abweichungsverstärker und einem Steuerteil. Sie können in Umgebungstemperaturen von 0...50° C und mit Netzanschluss betrieben werden.

Die Typen 18-92 und 18-98 haben Schalter mit sieben Drucktasten und ein 10gängiges Präzisionspotentiometer für die Sollwerteinstellung. Jede Taste gibt eine Stufe von 100°, und das Potentiometer unterteilt jede 100°-Stufe in 1000 Skalenteilungen, so dass sich eine Einstellbarkeit von 0,1° ergibt. Das Eingangssignal wird gefiltert und dann mit dem Ausgangssignal des Sollwertgebers verglichen. Das entstehende Fehlersignal wird in den Abweichungsverstärker gespeist, dessen Ausgang ein Abweichungsmessgerät treibt. Im Typ 18-92 wird der auf dem Abweichungsmessgerät angezeigte Ausgang in einen Steuerteil gespeist, der ein Gleichstromsignal von 0...5 mA abgibt. Der Steuerteil des 18-98 gibt Zündungsimpulse ab, die zum Treiben von Thyristoren geeignet sind.

Die Typen 18-99 bzw. 18-92 unterscheiden sich von den Typen 18-92 bzw. 18-98 nur in einer Beziehung: sie sind Differenz-Thermoelementwächter. Statt der digitalen Sollwerteinstellung haben sie geeichte Skalen mit 50 gleichen Unterteilungen. Die Standardeichung ist −0,5 mV...+0,5 mV.

EE 87 765 *für weitere Einzelheiten*

Niedertouriger Synchronmotor

Vactric Control Equipment Ltd, Garth Road, Morden, Surrey

(Abbildung Seite 766)

Die Fertigung einer neuen Ausführung des niedertourigen Synchronmotors "Vacsyn" läuft im Werk Morden der Vactric Control Equipment Ltd an; sie ist besonders für Anwendungsgebiete geeignet, in denen eine langsame, konstante Geschwindigkeit erforderlich ist, wie z.B. Bürokopiergeräte, Absperrschiebersteuerung oder Schaltmechanismus von Werkzeugmaschinen.

Zur Erleichterung kompakter Installation in beschränktem Raum wurde der Motordurchmesser auf weniger als 101 mm heruntergebracht, trotzdem die Originalleistung von 6,6 W beibehalten wurde. In der Standardausführung ist der Motor für 115 V 50 Hz gewickelt und hat ein Mindestdrehmoment von 10,8 cm kp; es sind Ausführungen mit 60 oder 71,4 UPM lieferbar.

Die niedrige Synchrondrehzahl des Motors wird durch einen dauernd magnetisierten mehrpoligen Läufer mit einem Stator, der eine symmetrische Zweiphasenwicklung hat, erreicht. Er ist für Betrieb am Einphasennetz ausgelegt, und die Phasenverschiebung für die zweite Wicklung erhält man durch externe Serienschaltung eines Widerstandes und Kondensators.

Man braucht keine teure Untersetzung, und der Motor ist selbsterregt und umkehrbar. Durch Anlegen einer Reihenfolge positiver und negativer Impulse an jede Wicklung kann er auch als Schrittschaltwerk Verwendung finden. Je nach Motortyp ruft jeder Polaritätswechsel eine Vorwärtsdrehung der Welle um 1,8° oder 2,14° hervor, so dass entweder 200 oder 168 Schritte eine Umdrehung vervollständigen.

EE 87 766 *für weitere Einzelheiten*

Mikroschalter

The Plessey Co. Ltd, Electromechanical Division, Havant, Hampshire

(Abbildung Seite 766)

Ein vom Geschäftsbereich Elektromechanik der Plessey-Gruppe Bauelemente herausgebrachter neuer Kleinstmikroschalter vereint hohe Empfindlichkeit mit mechanischer Langlebensdauer.

Der mit Typ 19 bezeichnete Schalter hat einen Messerkontaktmechanismus mit nur zwei beweglichen Teilen und benötigt für ultraschnelles und empfindliches Funktionieren nur eine Bewegungsdifferenz von 0,013 mm. Die mechanische Lebensdauer überschreitet 10 Millionen Vorgänge, während die elektrische bei voller Nennlast von bis zu 5 A, 250 V∼ 100 000 Vorgänge beträgt. Bei halber Nennlast ist die nominelle elektrische Lebensdauer 250 000 Vorgänge.

Grundsätzlich ist der Schalter ein einpoliger Umschalter; er kann jedoch auch als einpoliger Ausschalter mit Arbeits- oder Ruhekontakten geliefert werden. Anschlüsse gibt es für Löten und Klemmen; der Standardkontaktstoff ist Silber. Im Sonderauftrag sind Kontakte in vergoldetem Silber oder Spezialwerkstoffen erhältlich.

Die Standardbetätigungskraft für den Mikroschalter 19 ist 115...140 p, und Betätigungseinrichtungen sind in grosser Auswahl lieferbar. Für normale Verwendungszwecke wird das nur 6,35 × 19,8 × 9,1 mm grosse Gehäuse aus Phenolharz gepresst; es stehen jedoch auch Diallylkunststoffe für Temperaturen bis zu 200° C zur Verfügung.

EE 87 767 *für weitere Einzelheiten*

Drehschalter

Honeywell Controls Ltd, Brentford, Middlesex

(Abbildung Seite 766)

Honeywell Controls Ltd stellt einen neuen anpassungsfähigen Drehschalter 1RM vor, der die Funktionen von bis zu acht Kippschaltern ausführen kann.

Der kompakte Baustein nimmt auf der Frontplatte nur 41,3 × 36,8 mm in Anspruch. Er ist robust gebaut und mit hochbelastbaren Honeywell-Schaltern V3 bestückt. Diese Schalter lassen sich in jeder beliebigen Reihenfolge betätigen—es gibt bis zu 40 000 Variationen —und in demselben Gehäuse können für verschiedene Strompegel bemessene Schalter vorgesehen werden; für niedrige Stromstärken werden Goldkontakte, für 15-A-Kreise Silberkontakte geliefert. Die Schalter werden durch eine Nockenwelle betätigt, deren Konstruktion Präzisionsfolgeschalten während einer langen Lebensdauer mit manueller oder motorisch angetriebener Betätigung gewährleistet.

Ein Vorteil des neuen Bausteins liegt in der Schnappfunktion der Schalter V3, durch die in herkömmlichen Drehschaltern auftretendes Weisses Rauschen beseitigt wird. Der Baustein kann durch mit Klebstoff beschichtete Dichtungen und 0-Ringe auf der Frontplatte abgedichtet werden.

EE 87 768 *für weitere Einzelheiten*

Selektive Verstärker

Avon Communication and Electronics Ltd, 14-16 High Street, Christchurch, Hampshire

(Abbildung Seite 767)

Die vor kurzem von Avon Communication & Electronics Ltd vorgestellten preiswerten und vielseitigen abstimmbaren selektiven Verstärker der Serie 400 haben breite Anwendungsmöglichkeiten in der Industrie wie auch in der Fernmeldetechnik.

Die Verstärker sind über den Frequenzbereich 270...3700 Hz in vier Bändern durchstimmbar. Die Selektivität kann zwischen 5 und 20 Prozent eingeregelt werden; für die dem Durchlassband benachbarten Frequenzen wird gute Unterdrückung durch Verwendung von zwei gleichzeitig abgestimmten Kreisen, deren Dämpfungsverlauf sich 120 dB/Oktave nähert, gewährleistet.

Die aktiven Elemente sind transistorisiert und werden mit internen Batterien von etwa drei Monaten Lebensdauer betrieben. Die Verstärker sind kompakt und robust gebaut und für Abstimmzwecke mit einem 64-mm-Messinstrument ausgerüstet.

Bei Einsatz in Fernmeldenetzwerken wird die Leistungsfähigkeit der selektiven Verstärker weiterhin durch Anwendung einer nichtlinearen Endstufe oder Torschaltung verbessert, da durch Einführung eines Schwellen- oder Mindestsignalpegels im Ausgang eine nützliche Verbesserung des Rauschabstandes erzielt werden kann.

Ein Gerät dieser Art kann den Rauschabstand einer Fernmeldeverbindung um mindestens 20 dB verbessern.

Für Anwendungsmöglichkeiten im Labor, z.B. für Brückenabgleich, lassen sich lineare Endstufen einbauen. Selektive Verstärker verbessern die Nullpunktgenauigkeit wesentlich. Um Überlastung des Gehörs beim Brückenabgleich mit Kopfhörer zu vermeiden, können die Modelle mit Begrenzerschaltungen ausgerüstet werden.

EE 87 769 *für weitere Einzelheiten*

Drehwähler mit Schutzrohrkontakten

Alma Components Ltd, Park Road, Diss, Norfolk

(Abbildung Seite 767)

Dieser neue Drehwähler mit Schutzrohrkontakten ist klein, zuverlässig, anpassungsfähig und wesentlich billiger als andere elektronische Schalteinrichtungen. Die Abmessungen eines kompletten 10stufigen Wählers mit drei Ebenen sind nur 89 × 89 × 76 mm. Dieser Baustein braucht nur eine 24-V-Stromversorgung und schaltet einen Schritt vorwärts, wenn zwei Drähte des Wählers kurzgeschlossen werden; es sind keine zusätzlichen Schaltungen erforderlich.

Man kann Betrieb in umgekehrter Richtung vorsehen und automatischen Rücklauf zu jeder beliebigen vorgewählten Nummer einbauen. Die Einheiten können in Dekaden geschaltet werden, und auf Wunsch gibt es dann auch "Vortrag" von Null auf die nächste Einheit. Jede Stufenanzahl bis zu 100 kann mit bis zu sechs Schaltebenen geliefert werden. Statt der üblichen Arbeitskontakte sind auch Umschalt- oder Ruhekontakte zu haben.

Aus der Verwendung von Schutzrohrkontakten ergeben sich viele Vorteile, u.a. niedriger Leistungsbedarf (½ W), hohe Arbeitsgeschwindigkeit, volleingeschmolzene Kontakte, die weder Wartung noch Justieren erfordern; dazu lange Lebensdauer, getrennte Kontakte statt der Schaltung mit dem gemeinsamen Schleifer und niedriger Kontaktwiderstand.

EE 87 770 *für weitere Einzelheiten*

Elektronische Thermometer

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(Abbildung Seite 767)

Das elektronische Thermometer 166C ist das neuste aus dem Comark-Fertigungsprogramm für Thermoelement-Temperaturanzeiger.

Modell 166C für Verwendung mit Cromel-Alumel-Thermoelement und Zuleitung beliebigen Widerstandes gibt Direktanzeige der Temperatur ohne Tabellen von $-120°$ C...$+1100°$ C mit einem Auflösungsvermögen von einem Grad bei allen Temperaturen.

Automatische Vergleichsstellenkompensation, Festkörper-Chopperverstärker und Langlebensdauer-Quecksilberbatterien geben schnelle, genaue Messungen mit einem Reglerminimum.

Die Ausgangsspannung des Thermopaares wird in 100°C-Stufen durch eine Spannung reduziert, die einem gegen ein internes Bezugselement eichbaren Präzisionsspannungsteiler entnommen wird. Das Messgerät überbrückt das 100°-Intervall. Ausserdem gibt es einen Suchbereich 0...1000°C.

Dem Standardausgang für ein Registriergerät können 200 μA entnommen werden.

EE 87 771 *für weitere Einzelheiten*

Halbleiter-Tester

Vertrieb: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(Abbildung Seite 767)

Der von Sercel S.A.R.L. in Frankreich hergestellte "Diotestor" wurde hauptsächlich für das Testen von Halbleitern auf gedruckten Schaltungen am Ende des Fliessbandes entwickelt. Das Gerät findet fehlerhafte Dioden und Transistoren ohne Ablöten der Bauelemente.

Der Diotestor prüft Leitfähigkeit und Grenzbedingungen in pnp/npn-Transistoren und Dioden. Die Testmethode ist einfach und erfordert keine besonderen Fachkenntnisse.

Es gibt zwei Modelle, von denen eins eine 3-V-Standardtrockenbatterie benutzt und etwa 148 g wiegt, das andere mit einem Voltabloc-Akku ausgerüstet ist, der zum Aufladen nur ans Netz angeschlossen wird. Das letztere Modell wiegt etwa 310 g.

Der mit Batterien betriebene Diotestor hat ein unzerbrechliches Gehäuse aus Rilsan, ist zweckdienlich gestaltet und leicht, so dass er sich im Einsatz einfach und bequem handhaben lässt.

Als Zubehör sind ein Adapter mit sieben Anschlüssen in verschiedenen Abständen für Diodentests und eine Verlängerungsschnur mit Buchse und Krokodilklemmen zum Testen nicht eingebauter Halbleiter lieferbar.

EE 87 772 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Die Belastung von Mehrkanal-Nachrichtenverkehrssystemen mit weissem Rauschen von W. Oliver

Zusammenfassung des Beitrages auf Seite 714-717

In diesem Beitrag werden die jetzt international angewendeten Testverfahren für Mehrkanal-Nachrichtenverkehrssysteme besprochen. Es wird gezeigt, dass Gausssches Rauschen dieselben statistischen Eingenschaften hat wie Mehrkanalsignale; Methoden zur Umwandlung der durch Messungen von Rauschleistungsquotienten und Rauschabstand usw. erhaltenen Ergebnisse werden gegeben.

Schrotrauschen und das Messen der Ladung des Elektrons von E. Mathieson

Zusammenfassung des Beitrages auf Seite 718-721

Einzelheiten des versuchsweisen Messens von Schrotrauschen im Strom einer thermionischen Diode werden gegeben.

Aus diesen Messungen ergibt sich für die Ladung des Elektrons ein Wert von 1,62 $\times$ 10^{-19} Coulomb; die Versuchsunsicherheit dieses Ergebnisses ist etwa 5 Prozent. Die Rolle von Untersuchungen dieser Art als Laborversuch für Studenten wird besprochen und eine einfache Darstellung der diesbezüglichen Theorie gegeben.

Die Nichtlinearität von Festwiderständen von P. L. Kirby

Zusammenfassung des Beitrages auf Seite 722-726

Die Anwendung eines Spannungskoeffizienten für die Definition der Grösse der Nichtlinearität von Festwiderständen weist ernsthafte Mängel auf, und es wird eine alternative Methode vorgeschlagen, in der das Messen der durch die dritte Harmonische hervorgerufenen Verzerrung eine Rolle spielt. In einem einfachen Fall wird eine Ausgangsgrösse der dritten Harmonischen erzeugt, die der dritten Potenz der angelegten Grundspannung proportional ist; auf dieser Grundlage wird eine neue Einheit abgeleitet, die "Index der dritten Harmonischen" genannt wird. Eine Erklärung für den Ursprung der Nichtlinearität wird vorgeschlagen und ein Grund für ihre Korrelation mit dem Stromrauschen gegeben. Typische Indexwerte für Schichtwiderstände über einen breiten Wertbereich werden gegeben und weitere Anwendungsmöglichkeiten einschliesslich der Verwendung für die Ausdeutung der grundsätzlichen Eigenschaften von Widerständen, für Testpflichtenblätter und eine Auswahlmethode zur Gewährleistung hoher Zuverlässigkeit vorgeschlagen.

Ein 1-kW-Wechselrichter für Netzfrequenzbetrieb von W. D. Gilmour

Zusammenfassung des Beitrages auf Seite 727-731

Ein Wechselrichter, der bei genau bekannten Frequenzen zwischen 45 und 65 Hz bis zu 1 kW mit guter Konstanz der Ausgangsspannung abgibt, wird beschrieben. Bei Belastung bis zu 300 W ist die Ausgangsspannung sinusförmig und wird danach mit ansteigender Last fortschreitend mehr rechteckförmig. Ein niederspanniges Regelsystem findet Anwendung.

Galvanisch gekoppelte Schaltkreise mit Siliziumtransistoren von J. D. Martin

Zusammenfassung des Beitrages auf Seite 732-735

Die Anwendung galvanisch gekoppelter Schaltkreise mit Siliziumtransistoren ist zwar wohlbekannt, Entwurfsdaten stehen jedoch kaum zur Verfügung. Dieses Problem wird untersucht, wobei auf den Entwurf zuverlässiger Schaltungen mehr Wert gelegt wird, als auf solche mit grösster Verstärkung. Die Kollektorsättigungsspannung von Transistoren wird besprochen, und es werden Vorschläge gemacht, die eine Schätzung der Entwurfsparameter gestatten. Erfahrungen mit Schaltungen, die auf diese Weise entworfen wurden, haben die Zweckdienlichkeit dieses Verfahrens bestätigt.

Ein 400-Hz-Präzisions-Drehstromgenerator von D. J. Miles

Zusammenfassung des Beitrages auf Seite 736-739

Eine Drehstromversorgung mit konstanter Amplitude sowie Frequenz wurde als Energieversorgung für das Rad von Trägheits-Qualitätskreiseln entwickelt. Bei Normalbelastung und Eingangssapnnungsschwankungen wird eine Amplitudenstabilität innerhalb ±0,01 Prozent erreicht.

Verstärkerentwurf mit Hilfe eines Digitalrechners von J. R. Abrahams

Zusammenfassung des Beitrages auf Seite 740-745

Der Beitrag zeigt, wie ein kleiner Digitalrechner für die Ausführung von Entwurfberechnungen elektronischer Schaltungen eingesetzt werden kann. Die Berechnung eines mit Transistoren bestückten, zweistufigen abgestimmten Verstärkers wird nach gewisser Vereinfachung von Parametern als numerisches Beispiel benutzt. Die Möglichkeit, vergleichsweise Aufwandsberechnungen durchzuführen und damit eine wirtschaftlich optimale Ausführung zu erreichen, wird auch besprochen.

Ein Kettenverfahren zur Lösung mehrmaschiger Probleme von N. M. Morris

Zusammenfassung des Beitrages auf Seite 745-746

Ein Kettenverfahren zur Lösung mehrmaschiger Probleme wird beschrieben. Es stützt sich auf die Taylorsche Reihe und ist auf jedes lineare Gleich- oder Wechselstromnetzwerk sowie einige nichtlineare anwendbar. Es bietet einen zweckdienlichen alternativen Zugang zur Auflösung mehrmaschiger Netzwerke, wenn kein Simulator zur Verfügung steht.

Ein transistorisierter Quotientenverstärker, der nach dem Prinzip der Berechnung verstrichener Zeit arbeitet
von A. A. Cremin und M. C. Sexton

Zusammenfassung des Beitrages auf Seite 747-748

Eine beschriebene Analogschaltung ist zur Bestimmung des Quotienten von zwei zeitabhängigen Eingangssignalen geeignet. Nach der angewandten Methode werden Schaltungen für verstrichene Zeit dazu benutzt, Samplingfrequenzen zu erzeugen, die wesentlich höher liegen als die Eingangssignalfrequenzen, wodurch ein im wesentlichen kontinuierliches Ausgangssignal entsteht, das dem Eingangsquotienten proportional ist.

Negativer Ausgangswiderstand in Konstant-Stromversorgungen von J. Diamond

Zusammenfassung des Beitrages auf Seite 749-751

In einer Konstant-Stromversorgung ist manchmal ein negativer Ausgangswiderstand wünschenswert. Er kann zum Annulieren des Widerstandes von Zuleitung oder anderen Bauelementen in Serie mit der Last benutzt werden, wenn die übliche Lösung—Fernabtasten der Verbraucherspannung—nicht anwendbar ist.

Der Entwurf der Stromversorgung wird besprochen und gezeigt, dass unabhängig von der Ausgangsspannung ein konstanter negativer Ausgangswiderstand erzielt werden kann. Der Temperaturkoeffizient des Ausgangswiderstandes ist auch regelbar.

Die Bestimmung der Widerstandsmatrix komplizierter linearer Netzwerke von W. A. Crabbe

Zusammenfassung des Beitrages auf Seite 752-753

Das Verfahren zum Erlangen der Widerstandsmatrix eines linearen Netzwerkes nach Kron wird besprochen. Danach wird das Verfahren auf einen einfachen Dreipol und schliesslich auf einen zweistufigen Verstärker mit Brückenrückkopplung angewendet. Durch Andwendung der Widerstandsmatrix werden die Ergebnisse für gewisse, relativ komplizierte Netzwerkverknüpfungen vereinfacht. Das Verfahren nach Kron zum Erlangen dieser Matrix gibt einen formalisierten Zugang, der weniger Fehler zulässt.