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# Commentary

ONE of the many problems arising from the increasing air traffic across the Atlantic is the need to provide adequate air to ground communication links over the route.

Several solutions have been put forward which are being closely studied by the Ministry of Aviation, and one of these is the use of satellites.

There is, however, a degree of uncertainty already existing regarding the use of satellites for global communication purposes and it now appears that the Ministry of Aviation has rejected this latest proposal on the grounds that it would involve a vast re-equipment in aircraft and could not, in any case, be introduced in the short time available before supersonic aircraft are crossing the Atlantic. It seems obvious, too, that the cost would be out of all proportion.

The use of the conventional h.f. and l.f. bands is being explored but considerable attention is being given to the v.h.f. band which undoubtedly gives the most reliable service.

But the v.h.f. band is limited to virtually a line of sight path and with aircraft flying at 30 000ft the range is no more than about 300 miles.

To overcome this limitation of range, a very novel proposal has been put forward by the Ministry of Aviation, namely that a number of specially designed ocean platforms* installed with the appropriate v.h.f. equipment should be moored in precisely known positions across the Atlantic to extend the coverage of the existing coastal stations in Ireland and Newfoundland.

Already a consortium entitled Seastation Telecommunications Ltd has been set up, comprising Submarine Cables Ltd (owned jointly by Associated Electrical Industries Ltd and British Insulated Callender's Cables Ltd) and Cammell Laird Ltd, in order to carry out a design study in conjunction with the Ministry of Aviation.

The platforms, which will have a high order of stability and station-keeping, will be linked by a common submarine cable connected at each side of the Atlantic to the normal telephone systems of the air traffic control centres, so that should a break occur in the submarine cable, communication can be restored by means of the existing transatlantic cables.

The number, and therefore the cost, of these platforms has not yet been determined. Three, sited along the great circle between Ireland and Newfoundland, would, together with the coastal stations, give more or less complete east-west coverage while four platforms, two being sited north of the great circle and two south, would give practically 'solid' coverage over the whole of the North Atlantic for aircraft at a height of 30 000ft with the additional coverage provided by the coastal stations at Greenland and Iceland.

The platforms themselves are of unique design consisting of a submerged hull structure well below the surface of the sea in order to give the required degree of stability, and in the form of a vertical cylinder some 400ft long and 30ft diameter. On the top of the cylinder will be mounted a flat platform with a diameter of 100ft on which the various aerials can be installed and which can also serve as a landing deck for helicopters.

Preliminary estimates set the cost of a single platform at about £1 million, and that of a complete three platform system including initial survey work, cables, terminal equipment etc. at about £12 to £14 million with an annual running cost of £250 000.

The main purpose of the system is, of course, to provide a v.h.f. radio telephone service for civil aircraft flying across the North Atlantic but a number of other facilities could be included such as marine h.f. communications and military aircraft u.h.f. channels. And with adequate cable capacity it would be possible to accommodate still more facilities such as communication services for airline company traffic and 'ship to shore' telephony which would become a source of income to offset the costs of the system.

Providing no difficulties arise due to siting or interference, certain types of point-source radio-navigational devices such as the m.f. non-directional beacons, v.o.r., and d.m.e., could also be installed for an overall aircraft navigation system across the North Atlantic. Such an installation could be justified on the ground that it would up-date aircraft inertial navigation equipment, resolve ambiguities in the data obtained from long range radio navigational aids and finally check gross errors in the overall navigation system.

There is also the possibility of a limited primary radar surveillance system but what looks more attractive is a secondary radar surveillance which, with data transmitted over the submarine cable to the main air traffic control centres, would allow a general monitoring and position fixing service for the control centres on both sides of the Atlantic similar to that already existing on a number of main overland routes.

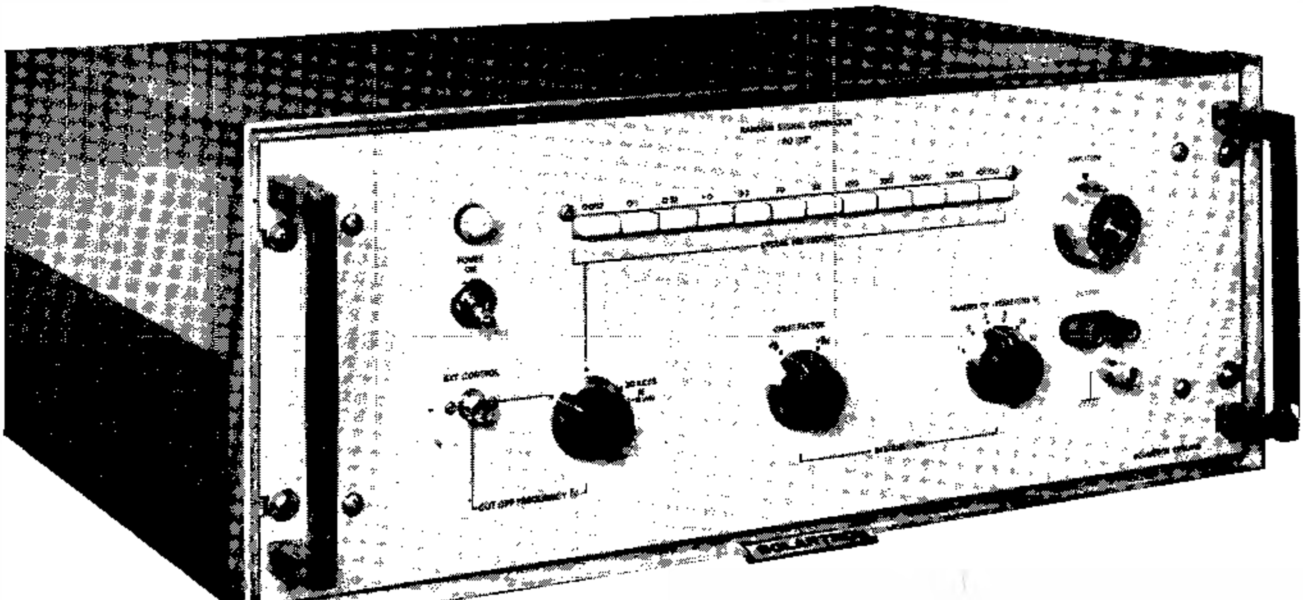
Still further use could, it seems, be made of the platforms and it has been suggested that they might well take over the function of the weather ships for the prompt and accurate transmission of weather data.

Altogether the proposed scheme appears to be thoroughly sound, and once the decision has been taken, it is anticipated that a three platform system could be brought into operation within three years.

If it proves successful, and this seems more than likely, there is every possibility that it could be expanded to a round-the-world system covering the main oceanic air routes.

* Seastations for Aircraft Communications. *Electronic Engng.* 37, 379 (1965).

# A Random Signal Generator



By D. A. G. Tait,* A.M.I.E.R.E., and M. Skinner*

Conventional noise sources have an output which can be regarded as being due to an infinity of infinitesimal events (e.g. ion collisions in a gas tube). A novel instrument is described wherein the signal is synthesized from a finite and controlled number of sources of well defined statistical parameters which results in a signal of unusual flexibility and stationarity.

(Voir page 62 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 69)

THE many ways of testing an electrical or mechanical system all involve the application of a stimulus and measurement of response, but differ in the type of stimulus and the characteristics of the response measuring device.

The classic approach in the measurement of electrical systems has been to apply a stimulus in the form of a nearly pure sine wave of voltage or current of adjustable amplitude and frequency, measurement of response being made (desirably) on an r.m.s.-reading voltmeter, though practically always on an average-measuring voltmeter. Amplitude is usually set to the maximum level that the system is known to be capable of handling without distortion, and measurement of output is made at various discrete input frequencies to yield a plot of amplitude versus frequency. Unfortunately it is found that this 'frequency-response' is very much dependent on the amplitude of applied signal except in highly linear systems operating at a conservative amplitude. For the same reason any measure of non-linearity (e.g. the inter-modulation distortion factor) is found to be dependent on the frequency used. Now it does sometimes happen that a system is required to work at only one frequency, or only one amplitude level, or both, but more usually it has been designed to cope with a variety of frequencies and amplitudes existing simultaneously—i.e. a complex waveform. It is therefore becoming increasingly common to test systems with a complex waveform, using, as a measure of the systems goodness, the mean square of the difference between input and suitably normalized output. This mean square error takes into account both amplitude—and frequency response errors over the range of amplitude and frequency covered by the test signal, and, in order to make these parameters representative of a wide variety of actual inputs, it is customary to use, as a test signal, a random waveform.

**Noise as a Test Signal**

A most important random signal is that of so-called Johnson noise, i.e. the noise voltage caused by thermal agitation in resistors (or in the resistive component of any impedance). This signal is characterized by a Gaussian probability distribution of instantaneous amplitudes, and a spectral density which is flat from zero frequency upwards, the high frequency limit being set only by stray capacitance, or by the bandwidth limitation of any measuring equipment.

An important consequence of these characteristics is that if the signal is passed through a band filter, the band restricted output will still have a Gaussian amplitude distribution. The power is of course reduced—in fact the power is directly proportional to bandwidth, this following from the fact that the original signal has a flat spectral density.

The Gaussian probability curve (Fig. 1 and Table 1) shows that extremely large amplitudes (in terms of the root-mean-square level) have a very small but finite, probability of being exceeded. It follows then, that any amplifier required to increase the level of such a signal without altering the amplitude distribution must needs have an infinite power handling capacity.

In practice it is usually accepted that the amplifier is sufficiently good if it will handle, without distortion, signals of peak amplitude equal to three times the r.m.s. amplitude (nine times the average power) since this peak level is

* *Solartron Electronic Group.*

only exceeded about 0·3 per cent of the time (see Table 1). However, note the considerable improvement given by an 80 per cent increase in power capacity (sigma = 4, $f(\sigma)$ = 0·006 per cent).

In most practical situations, the signal going into a system will have a defined peak amplitude, since it will be derived from a source which itself has a defined signal handling capacity. For instance, the a.f. amplifier following a radio tuner unit will not be called on to handle signals equivalent to more than 100 per cent modulation of the (controlled amplitude) carrier. It appears then that testing of such an amplifier would best be done using as a test signal one having a defined peak amplitude. It might be assumed that in such cases the 'sine-wave' signal would be best suited, but this is not so on account of the peak to r.m.s. ratio being unfavourable, resulting often in overheating of output stages when they are run at near full power. Often, too, bias conditions are altered by this 'crest-factor' of the test signal, which can cause wide discrepancies between measurements made under test using sine-waves, and observed results using real programme signals.

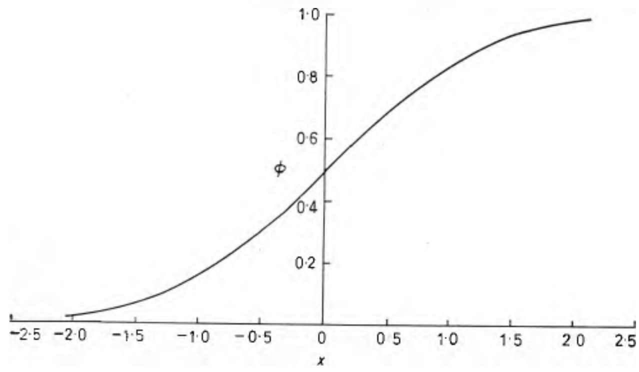
It was with these and similar arguments in mind that the present random signal generator was conceived, which enables the modern development or test engineer to make measurements on systems using a random signal of defined but adjustable crest factor (i.e. peak to r.m.s. ratio). The output power is sufficiently well defined as to render quite unnecessary the provision of a meter. This is a boon when operating at very low frequencies (< 10c/s) since

**Fig. 1(a). Gaussian probability distribution**

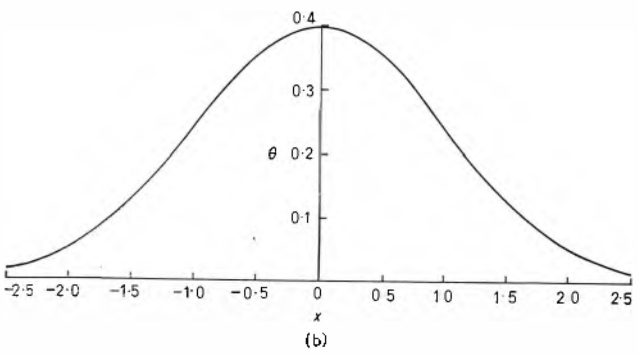
$$\phi(v) = \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{v} \exp(-\tfrac{1}{2}x^2)dx$$

**(b). Density function corresponding to Fig. 1(a)**

$$\theta = \frac{1}{\sqrt{2\pi}} \exp(-\tfrac{1}{2}x^2)$$




(a)




(b)

**TABLE 1**

**P is the probability (expressed as a percentage) that a normally distributed function of zero mean value will exceed (in magnitude) X times the r.m.s. value.**

| $P(\%)$ | $X(\delta)$ | $P(\%)$ | $X(\delta)$ |
|---|---|---|---|
| 100 | 0 | 1·0 | 2·58 |
| 50 | 0·675 | 0·27 | 3·0 |
| 37·1 | 1·0 | 0·1 | 3·3 |
| 20 | 1·28 | 0·01 | 3·9 |
| 10 | 1·65 | 0·006 | 4·0 |
| 5·55 | 2·0 | | |

the randomness of the output makes measurement of power a very tedious business. A further feature of this instrument is that the bandwidth is readily adjusted without altering the total output power of the system. This allows the full power of the system to be developed in a bandwidth of 0 to 1c/s or less for (say) servo-testing, or to be spread over the whole a.f. range 0 to 16kc/s or more, for the testing of amplifiers and communication channels. This bandwidth adjustment facility is also of great value in vibration testing, where it is usually desired to shake a test-bed with the maximum possible power in a given bandwidth.

### Target Specification of the Random Signal Generator

OUTPUT AMPLITUDE

Well defined at 10V peak-to-peak through 600Ω, level independent of supply and temperature variations or the setting of any controls.

SPECTRAL DENSITY

Flat down to zero frequency. High frequency response to follow law $V(f) = V(o) \dfrac{\sin \pi f/f_c}{\pi f/f_c}$ with cut-off frequency adjustable in semi-decade steps from 0·032c/s (≃ 2 cycles per minute) to 10kc/s. An additional range available at 20kc/s. Other cut-off frequencies to be achieved by external control.

CREST FACTOR

The peak-to-r.m.s. ratio adjustable in 12 steps from unity (random 'telegraps signal') to nearly 10 (near-Gaussian wave of 128 discrete levels) without change of peak level.

AMPLITUDE DISTRIBUTION

The waveform made up of discrete levels with probabilities determined by the 'Distribution' switch. The following probability density functions to be available among others.

(1) Rectangular (i.e. equal probability all levels) with up to 128 discrete levels. This is quivalent to a random number generator.

(2) Triangular

(3) Near-Gaussian with crest factor equal to 9·8.

DIGITAL OUTPUT

Available at rear of instrument.

LEVEL ACCURACY

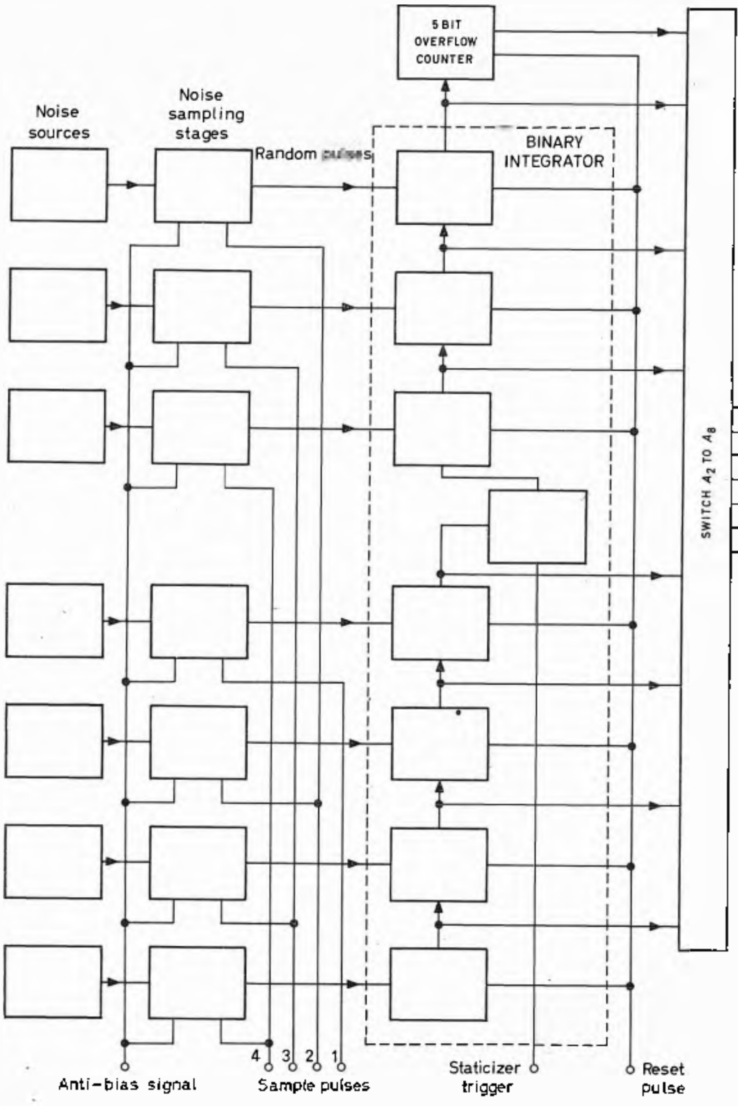
All discrete levels, and the mean level, defined to 0·2 per cent.

STATISTICAL ACCURACY

R.M.S. level and discrete level probabilities defined to 2 per cent.

### Principle of Operation

The principle is that of the 'Central Limit' theorem, which, put into electronics language, states that if a sufficient number of independent random sources contribute each a very small amount to the output, then the



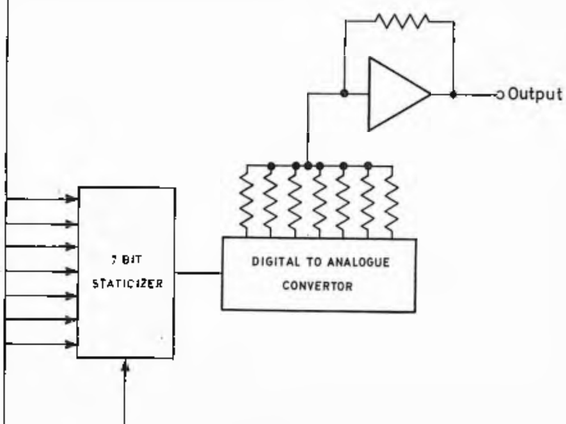



**Fig. 2. Signal circuit arrangement**

$$P({}_n{}^k) = \frac{n!}{k!\,(n-k)!}\,.\,P(1)^k\,.\,P(0)^{n-k}$$

$$= \frac{n!}{k!\,(n-k)!}\,.\,(\tfrac{1}{2})^n$$

output amplitude distribution will be Gaussian, regardless of the distribution of the individual sources. By using a variable number of independent signals each of equal and well-defined amplitude, an output signal is produced whose distribution is variable, and whose peak amplitude is well-defined, yet which satisfies the criterion of randomness.

The multiplicity of independent sources is achieved in two ways—by using seven separate noise tubes, and by sampling their outputs regularly at time intervals long compared with the reciprocal of their frequency bandwidth. Refer to the signal circuit schematic, Fig. 2.

The noise output of each source is sampled at regular intervals. The noise sampling stage effectively decides whether the instantaneous noise voltage is positive or negative and accordingly generates a pulse at one or other of its two output terminals. The presence or absence of an output pulse can be given logical 'one' or 'nought' nomenclature, then, provided that the noise power is large compared with any inherent bias in the sampling stage, the probability of presence or absence of a pulse at one of the output terminals is equal, i.e.,

$$P(1) = P(0) = \tfrac{1}{2}$$

Now the probability of an output being present $K$ times in a sequence of $n$ samples is given by the binomial distribution.

The output of each random pulse source is fed into the appropriate stage of a digital integrator for $n$ successive samples. The output of the integrator at the end of this period is transferred to a staticizer, this being connected to a digital to analogue convertor. The output of this is the system output for the next interval. The probability distribution of the amplitude of this output voltage, for the case of a single pulse source, is given by the expression:

$$f(v) = \sum_{k=0}^{(v)} \frac{n!}{k!\,(n-k)!}\,.\,1/2^n$$

$$= n!/2^n \left[\frac{1}{0!\,(n)\,(n-1)\ldots(1)} + \frac{1}{1!\,(n-1)\,(n-2)\ldots(1)} \ldots\ldots + \frac{1}{v!\,(n-v)\ldots(1)}\right.$$

$$= 1/2^n\left[1/0! + (n/1!) + \frac{n(n-1)}{2!}\ldots\frac{n(n-1)\ldots(n-v+1)}{v!}\right.$$

The series within the brackets can be written down by inspection—it is the $n^{th}$ line of Pascal's triangle thus:

```
          1   1
        1   2   1
      1   3   3   1
    1   4   6   4   1
  1   5  10  10   5   1
1   6   etc.
```

In fact seven random pulse sources feed the successive stages of the digital integrator (counter). The result of this is that instead of having (for $n = 1$) only 2 levels each of probability $\frac{1}{2}$ there are 128 ($2^7$) equally probable levels. In this case the series may be written down along the following lines (which example, through lack of space, is confined to the case of 2 pulse sources only).

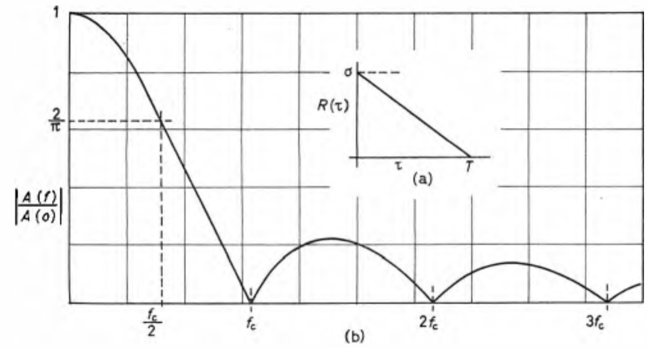



**Fig. 3(a). Auto-correlation function $T = r/f_c$**
**(b). Normalized amplitude spectral density**

| | | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | 0 | 0 | 0 | 1 | 1 | 1 | 1 | 0 | 0 | 0 | | $n$ = 1 |
| | | 0 | 0 | 1 | 2 | 3 | 4 | 3 | 2 | 1 | 0 | 0 | | 2 |
| | 0 | 1 | 3 | 6 | 10 | 12 | 12 | 10 | 6 | 3 | 1 | 0 | | 3 |
| 1 | 4 | 10 | 20 | 31 | 40 | 44 | 40 | 31 | 20 | 10 | 4 | 1 | | 4 |
| 1 | 5 | 15 | 35 | 65 | 101 | etc. | | | | | | | | 5 |

In this trapezium each number is derived by the addition of the four numbers symmetrically disposed in the preceding line. Note that the distribution in the first line will be rectangular, and in the second, triangular, regardless of the number of primary levels.

If six of the random pulse sources are switched out, the output for $n = 1$ is a random telegraph signal, and the full output can be regarded as having been synthesized from a controlled number of uncorrelated but statistically identical telegraph signals. Synthesis of such signals differing in amplitude by a binary weighting results in a signal having the same shape of amplitude distribution as its constituents, but having many more discrete levels. Synthesis by signals having equal weightings results in the 'centralizing' tendency of the resultant amplitude distribution.

### Peak-to-R.M.S. Ratio

On the basis of each level of output amplitude being proportional to the sum of $n$ independent samples of the same source one may argue as follows.

The peak output is obtained when all the samples have their maximum value (assumed to be +1 unit). Then the peak output amplitude is proportional to $n$, therefore the peak power is proportional to $n^2$.

Now the average power output is proportional to $n$ times average power of each sample, on the assumption of zero correlation. The peak to average power ratio is therefore: $n^2/nK = n/K$ where $K$ is the ratio of average-to-peak power of single samples.

If each sample has only 2 possible values (+1 or −1 units) its peak-to-average power ratio is unity, therefore the integrated output has a peak-to-average-power ratio $n$ times, therefore the crest factor (peak amplitude to r.m.s. amplitude) = $\sqrt{n}$. If each sample has 128 possible levels with equal increments and all equally probable, its peak to average power ratio is very nearly 3 times, giving an integrated crest factor of $\sqrt{3n}$. This is shown more formally in the Appendix.

### Power Spectrum

This is a sampled system, and the output waveform holds at a constant level between successive sampling instants. When there is zero correlation between successive pulses, the auto-correlation function is as shown in Fig. 3(a), and the corresponding spectral density as in 3(b).

### D.C. Offset

Whereas it was regarded as reasonable to control the statistical properties of the signal to about 2 per cent, it was decided that the mean level of the output should be controlled to 0·2 per cent. Residual bias in the random pulse sources is eliminated by periodically reversing their output connexions. This is discussed further in the circuit description.

## Circuit Description

The circuit may be divided into two main sections:
(a) the control system and (b) the signal circuits.

The control system, (see Fig. 4), generates a set of signals which govern the system. It is an open loop and each set of signals is initiated by a trigger pulse from the master oscillator.

The master oscillator uses an emitter-coupled circuit, which may be switched to any one of the number of preset frequencies; these range in semi-decade steps between 10kc/s and 0·032c/s, an additional spot frequency of 20kc/s is also available. In order to achieve the very low frequencies, and give economy of components, a decade counter is switched in series with the oscillator on certain ranges.

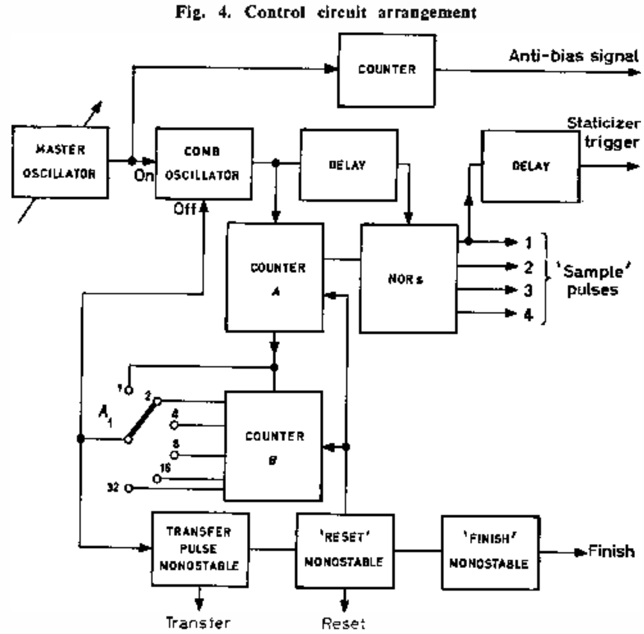
The output of the master oscillator is used to trigger a comb oscillator. This is a switched emitter-coupled oscillator, preset to a frequency of 1·4Mc/s. This oscillator is switched on by the output of the master oscillator and off by the output of switch $A_1$.

The output of the oscillator will comprise of a train of pulses the length of which will be defined by the position of $A_1$, which can be set to the output of counter $A$ or any of the stages of counter $B$. Counter $A$ comprises of two bits and counter $B$ of five bits, consequently the length of the pulse train will vary between 4 and 128 pulses.

The output of switch $A$ is also used to trigger a set of three cascaded monostable circuits; these generate a transfer pulse, reset pulse and a cycle finish pulse respectively. The transfer and reset pulses are essential to the operation of these signal circuits, while the finish pulse is available with the digital output. An additional output of the reset pulse generator is used to reset counters $A$ and $B$ to the start conditions.

Noise circuit sample pulses are derived from a set of four NOR circuits. They are set to detect each state of counter $A$; a pulse from each NOR is generated for each cycle of counter $A$, and are timed by a delayed version of the output of the comb oscillator. The trailing edge of number one sample pulse is differentiated to generate a

**Fig. 4. Control circuit arrangement**



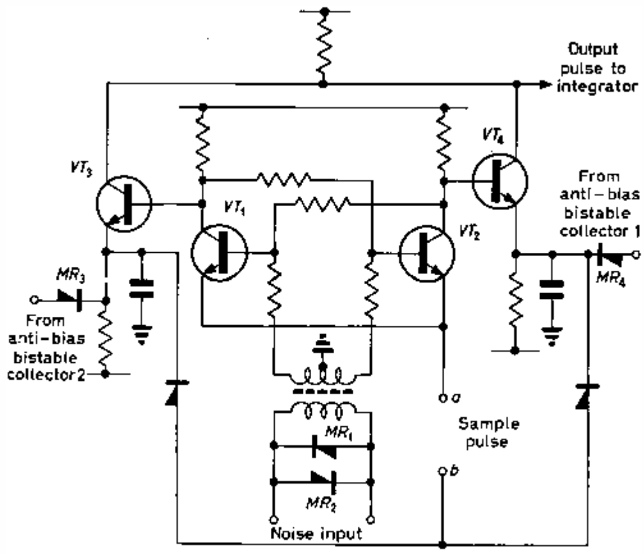



**Fig. 5. Noise sampling and anti-bias circuit**

further pulse which is used to trigger a staticizing stage in the digital integrator.

Finally, a separate counter stage is triggered by the output of the master oscillator. The output of this counter is used to remove bias in the noise sampling circuits.

The signals generated by the control circuits are used to control the operation of the random signal generating circuits. The signals may be summarized as follows:

(a) A set of four interlaced pulse trains which govern the operation of the noise sampling circuits.
(b) A half frequency square wave to remove zero frequency bias from the output.
(c) A trigger pulse to determine the state of the accumulator staticizer.
(d) A transfer pulse, whose function is to transfer binary information from the integrator to the output staticizer.
(e) A reset pulse to reset all counter stages to the start condition.
(f) A 'cycle complete' pulse.

The signal generating circuits (see Fig. 2) operate with random signals derived from the noise sources, and provide both analogue and digital outputs.

The seven noise sources are independent, as are the sampling circuits which follow; the output of the sampling circuits are assigned a binary weighting and drive the appropriate stage of a binary integrator.

Each of the seven noise sources consists of a thyratron with an annular magnet to eliminate spurious oscillations. The noise sources give substantially white noise over a band of about 1Mc/s, with an amplitude of 3V r.m.s. This signal is squared off at 0·5V peak-to-peak (see Fig. 5) by the back-to-back diode pair $MR_1$ and $MR_2$ and transformer coupled to the bistable pair $VT_1$ and $VT_2$. Application of a noise sampling pulse to point (*a*) causes the bistable to take the polarity of the noise appearing across the transformer output.

Hysteresis in the bistable is avoided by the sampling process since the 'bistable' is switched off during the period between samplings, allowing time for both bases to recover to their normal unbiased state. An alternative viewpoint is that the transistor feedback pair forms a super-regenerative pair, whose regenerative gain is a function of the emitter current controlled by the sampling pulse. The decision as to which way the bistable will tumble is made during the rise-time of the sampling pulse.

$VT_3$ and $VT_4$ comprise the anti-bias circuit; application of anti-phase signals to the emitters via $MR_3$ and $MR_4$ will select one transistor of the pair, as the operative side. Provided that the selected side of the noise bistable does not conduct on the application of the sampling pulse, an output pulse will appear at the summed collectors of $VT_3$ and $VT_4$. Obviously satisfactory operation of this circuit will only occur if the sample pulse applied to points *a* and *b* appear at the same instant in time but are shifted in d.c. level.

One stage of the integrator is shown in Fig. 6. The integrator consists of seven cascaded counter stages, with independent input to each bit of the counter from a noise sampling circuit. The integrator chain is broken after the fourth stage by a staticizer stage. The output of this stage is connected to the input of the next bit of counter.

It will be appreciated from the description of the control system that the noise sampling pulses consist of four interlaced pulse trains. The length of these pulse trains is defined by the position of switch *A* and vary in length from one pulse, in binary increments, to 32 pulses. Obviously, when a pulse train of length greater than one is applied to the noise sampling circuit, a random pulse train will emerge. The average number of pulses in this train will be half that of the sampling pulse train.

Application of this random pulse train to the noise input of a counter stage will cause the stage to count, and to produce carry inputs to the next stage of the cascaded integration chain. As the sample pulses are interlaced this carry will appear at a time not occupied by a noise pulse.

As the integrator has seven inputs and is seven stages long, five further counter stages are cascaded with the integrators to accept the overflow when 32 integration cycles are demanded.

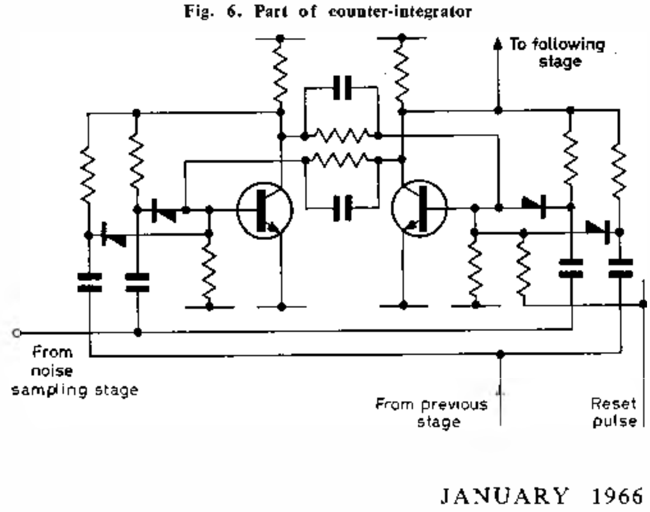
The output stage of this noise generating system consists of a seven bit staticizer and a precision digital to analogue convertor.

The staticizer comprises of seven bistable stages. The information regarding the state of the staticizer is transferred from the integrator via switch $A_{2-7}$ by means of the transfer pulse which occurs at the end of the integration cycle. The integrator and overflow counter are then reset to zero.

The group of seven stages of the integrator system from which the staticizer information is derived, is defined by the length of the integration cycle. For one integration the 7 least significant stages are selected; for 2 integrations the least significant bit is dropped, and the next 7 stages selected. The position of the staticizer is advanced one further bit, along the integrator, each time the order of integration is advanced.

Buffer stages are connected to the staticizer and provide the binary output of the instrument. The analogue output

**Fig. 6. Part of counter-integrator**



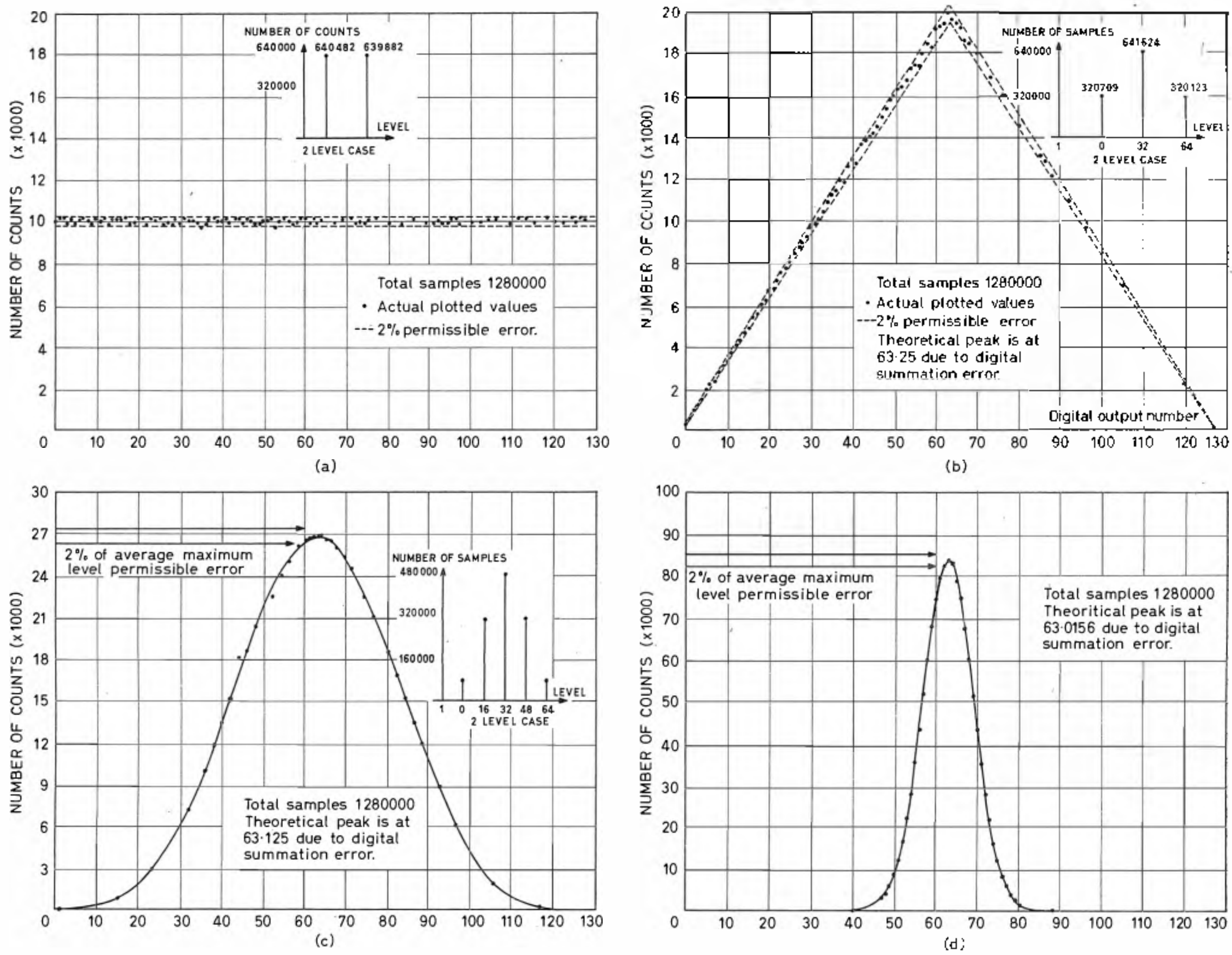



**Fig. 7(a). Measured distribution (rectangular) $N=1$** **(b). Measured distribution (triangular) $N=2$**
**(c). Measured distribution, $N=4$** **(d). Measured distribution, $N=32$**

is derived from a chopper-stabilized operational amplifier. A parallel transfer of information from the staticizer to the input of the amplifier is provided by binary weighted resistors. The gain of the amplifier is defined by the ratio of the feedback resistor to the input resistors. The adoption of this system provides a low impedance output with a precisely defined level, which may be held over any period of time.

### Results

One of the by-products of the digital synthesis technique is that measurements of the amplitude distribution are readily made by coupling the digital output to a core-store or other counter. Fig. 7 shows the distributions thus measured. Since the digital output is restricted to seven bits, there is a 'rounding-off' error in the later (larger $n$) distributions. This results in the probabilities ideally associated with the missing levels appearing at their adjacent lower level, thus displacing the distribution curve by up to one half (actual) bit. This represents a shift in mean level which has to be removed in the digital to analogue conversion.

### Conclusion

Because of the defined (binomial) amplitude distribution and defined spectral density it is believed that this instrument will be of value even outside those fields for which it was conceived. For example, the probability of any level being achieved can be readily estimated, which should be useful in fatigue work (stress/strain analysis). Again, because the changes in output level can be slowed down indefinitely, it should have considerable utility in physiological and psychological experiments. Its high degree of stationarity may be of greatest value in computer work, and the defined spectral density is important in environmental testing ('shaking').

### Appendix

The counter-chain repetitively performs the sum

$$Y = X_1 + X_2 \dots + X_N$$

where the $X_N$ are the $N$ sequential randomly-triggered operations of bistable stages.

Taking means

$$\langle Y \rangle = \langle X_1 \rangle + \langle X_2 \rangle \dots + \langle X_N \rangle$$
$$= 0 \text{ (by arrangement of bias, etc.).}$$

Taking the square

$$Y^2 = X_1^2 + X_2^2 \dots + X_N^2 + \text{ cross-products.}$$

Taking the mean of the square

$$\langle Y^2 \rangle = \langle X_1^2 \rangle + \langle X_2^2 \rangle \dots + \langle X_N^2 \rangle + 0,$$

since the mean value of all cross-products is zero (zero correlation).

But $\langle X_1^2 \rangle = \langle X_2^2 \rangle = \langle X_N^2 \rangle$ etc, because all come from the same source.

$$\therefore \langle Y^2 \rangle = N \langle X^2 \rangle$$
$$\therefore \sqrt{\langle Y^2 \rangle} = \sqrt{N} \sqrt{\langle X^2 \rangle}$$

But the peak value of the output $Y_{pk} = NX_{pk}$,

$$\therefore \text{ the Crest Factor} = \frac{Y_{pk}}{\sqrt{\langle Y^2 \rangle}} = \sqrt{N} \frac{X_{pk}}{\sqrt{\langle X^2 \rangle}}$$

i.e., the Crest Factor of the sum of the $N$ sequential inputs to the counter equals $\sqrt{N}$ times the Crest Factor of each input.

# A Wideband Transistorized Power Amplifier

By R. C. French*, A.M.I.E.R.E.

*A wideband transistorized power amplifier is described which has a gain of 27dB, a 3dB bandwidth of 2Mc/s to 60Mc/s and an output swing of 22V peak-to-peak into 75Ω at 30Mc/s. The output power is 0·8W.*

(Voir page 62 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 69)

A WIDEBAND transistor power amplifier for driving ultrasonic quartz delay lines is described. The delay lines in use normally have a loss between 20dB and 60dB. In order to conserve signal-to-noise ratio the drive amplifier should give as high an output power as possible consistent with adequate bandwidth, gain and linearity. The use of loop feedback over several stages avoids the reduction in output power caused by single stage feedback. In single stage feedback a large proportion of the output power is absorbed in the feedback network due to the low input impedance of power transistors. The much greater bandwidth achieved in small signal amplifiers cannot be obtained at power levels of one watt due to the low transition frequency of available power transistors. The amplifier is built on a printed circuit. It was found to perform satisfactorily as a delay line driver, and to offer a considerable advantage over a valve equivalent in terms of physical size and ease of construction.

### Choice of Transistors

The Mullard BFY70, a silicon planar epitaxial transistor, is used in the power outut stage. The BFY70 has a transition frequency $f_T = 210$Mc/s and a collector dissipation $P_c = 5$W. The AFY19 is used in the input and driver stages and has an $f_T = 350$Mc/s and $P_c = 0{\cdot}8$W.

### Possible Configurations

The choice between common emitter and common base must be made. The greater bandwidth and higher linearity of the common base arrangement are outweighed in this application because the transistor will be used up to frequencies three or four times greater than $f_{hfe}$ where $f_{hfe} = f_T/h_{fe}$. At frequencies above $f_{hfe}$ the current gain is dropping at 6dB/octave. At, say, $4f_{hfe}$ the current gain is 0·25 and the current swing which must be supplied to the base of the transistor is four times the current swing in the load resistor. The driver transistor must therefore be a higher current transistor than the output transistor. To obtain a high collector current a high collector dissipation transistor has to be used, so that the driver transistor has to be as large if not larger than the output transistor. This anomaly can be avoided if a step-down transformer is used. However, a wideband transformer having a large turns ratio and able to pass up to 0·5W would prove difficult. The common emitter configuration is therefore used for the output stage and by similar reasoning for the driver stages as well.

### Correction of Frequency Response

#### Single Stage Correction

Since the amplifier is to be used at frequencies well above $f_{hfe}$ some form of correction must be used to obtain a level gain-frequency response. The correction may be passive and of the form shown in Fig. 1. Here the time-constant of the inductor $L_c$ and load $R_L$ is made the reciprocal of $f_{hfe}$, and a fair measure of correction can be obtained. However, any change in $f_{hfe}$ with time or change in transistor will upset the frequency response. A more reliable method of correction using feedback is shown in Fig. 2 where both shunt and series feedback are illustrated. The inductor $L$ is used to supply d.c. power, not to correct the frequency response. The application of

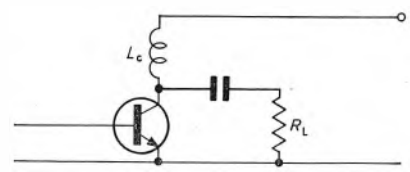
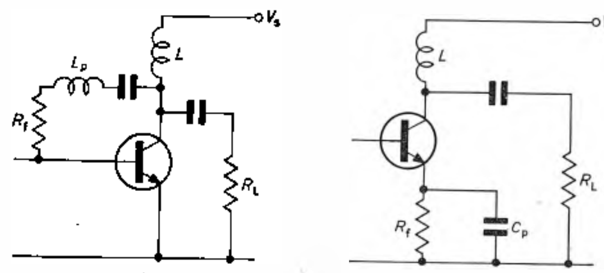



**Fig. 1. Compensated stage**




(a) (b)

**Fig. 2(a). Shunt feedback**
**(b). Series feedback**

feedback produces a very satisfactory frequency response but absorbs a large proportion of the output power at low frequencies where the 'peaking' element, $L_p$ or $C_p$, has little effect. In the case of shunt feedback the resistor $R_f$ must, to provide sufficient corrective feedback, be of the same order as the input impedance of the transistor. The input impedance is usually less than the load impedance so that a large proportion of the output current is lost into the feedback circuit, resulting in a considerable reduction in output power. An equivalent effect occurs with series derived feedback where the possible collector swing is reduced by the voltage swing at the emitter due to the undecoupled emitter resistor.

#### Multi-stage Correction

The loss in available power output can be considerably reduced by applying feedback over more than one stage. The value of $R_f$ is increased in the case of shunt feedback, or reduced in the case of series feedback, by the greater loop gain.

The only practical arrangement for a two stage circuit is shown in Fig. 3. The increase in loop gain tends to increase the value of $R_f$ required, however this is partially offset by the low impedance seen at the emitter of $VT_1$. This low impedance is the parallel combination of the resistor $R_e$ and the low impedance seen looking into the emitter of $VT_1$. $R_e$ must be small to avoid excessive feed-

* *Mullard Research Laboratories.*

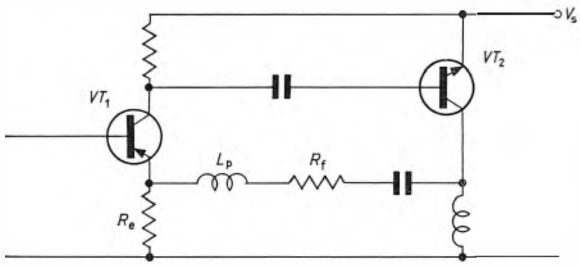



**Fig. 3. Feedback double**

back in $VT_1$ and consequent loss of available power to drive the output transistor. The improvement over single stage feedback is therefore marginal and the loss in output power unacceptable.

The next possibility is a three-stage amplifier using common emitter configurations with loop feedback which may be either shunt or series derived. Shunt feedback was investigated first and found to be satisfactory, although there is no reason why series feedback should not be just as effective.

### Three Stage Amplifier

The circuit arrangement is shown in Fig. 4. The first two stages have emitter feedback with feedback capacitors, the use of individual feedback for the first two stages being necessary to maintain stability. Without this feedback the loop would contain four lags having very similar time-constants and the amplifier would oscillate if sufficient loop feedback were applied. In the circuit shown, the first two stages are designed to have constant gain up to 50Mc/s so that the loop feedback is only required to correct the gain of the output transistor. Further, the loop feedback is designed to be small at the high frequency end of the band ensuring a high margin of stability.

### D.C. Design

OUTPUT STAGE

The output transistor used is the Mullard **BFY70**. Measurements made show that the thermal conductivity with an 11in² heat sink of aluminium is:

$$\theta_{ja} = 37°C/W$$

With a maximum junction temperature of 175°C and a maximum ambient of 33°C the maximum dissipation is given by:

$$P_c = \frac{175 - 33}{37} = 3{\cdot}85W$$

A suitable load resistor for maximum power output must have such a value that at full collector dissipation neither the maximum collector current nor the maximum collector emitter voltage is exceeded.

Now:

$$V_{ceq} = I_{cq}R_L + V_{ce(sat)}$$

where $V_{ceq}$ = Quiescent collector emitter voltage

$I_{cq}$ = Quiescent collector current

= Peak signal current swing,

and $V_{ce(sat)}$ = Collector emitter bottoming voltage.

$V_{ce(sat)}$ is approximately 3·0V and a suitable load would be 75Ω.

When $V_{ceq} = 75\,I_{cq} + 3$

$P_c = 75\,I_{cq}^2 + 3\,I_{cq}$

where $P_c$ is the collector dissipation, equal to 3·85W.

$$\therefore\ 75\,I_{cq}^2 + 3\,I_{cq} - 3{\cdot}85 = 0$$

and $I_{cq} = 207\text{mA}$

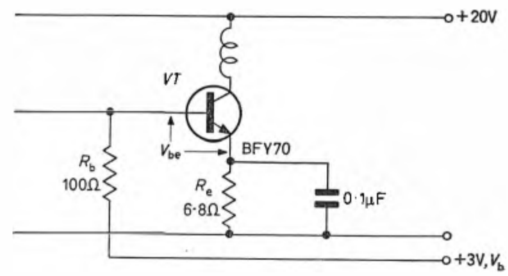



**Fig. 5. Output stage**

and maximum value of $V_{ceq}$ is:

$$V_{ceq} = P_c/I_{cq}$$
$$= 18{\cdot}6\text{V}$$

Now $I_{c(max)} = 2I_{cq} = 0{\cdot}42\text{A}$

and $V_{ce(max)} = 2\,I_{cq}R_L + 3$

$= 34\text{V}.$

Both collector current and voltage are safely less than the maximum values of 1·0A and 66V.

Variations in collector current due to changes in ambient temperature are, for the BFY70, almost entirely caused by variations in $V_{be}$.

Consider the circuit of the output stage shown in Fig. 5.

$$V_b = I_bR_b + I_eR_e + V_{be}$$

Taking small changes

$$0 = \delta I_bR_b + \delta I_eR_e + \delta V_{be}$$

when

$$\delta I_cR_b/\alpha' + (\delta I_c/\alpha)\,R_e = -\delta V_{be}$$

$$\therefore\ \delta I_c = \frac{-\delta V_{be}\,\alpha\alpha'}{\alpha R_b + \alpha' R}$$

$$\simeq \frac{-\alpha'\,\delta V_{be}}{R_b + \alpha' R_e}$$

now $R_e = 6{\cdot}8\Omega$

$R_b = 100\Omega$

$\alpha' = 4$

and $V_{be}$ decreases 2mV/°C.

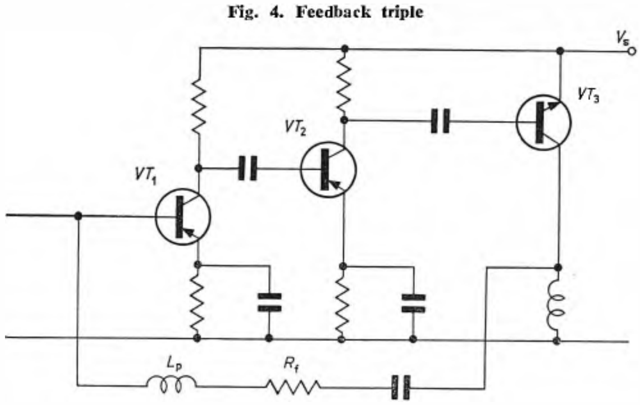
Consider the temperature range 10°C to 30°C

$$\delta V_{be} = -40\text{mV}$$

$$\delta I_c = \frac{40 \times 4}{10^3\,(100 + 27{\cdot}2)}$$

$$= 1{\cdot}25\text{mA}.$$

**Fig. 4. Feedback triple**



In order to determine the requirements for the driver stage, the gain and input resistance were measured and found to be ×20 and 17Ω respectively. The required input voltage for an output of 25V peak-to-peak is 1·25V peak-to-peak.

### Driver Stage

The required output of 1·25V peak-to-peak into 17Ω requires a current swing in the driver transistor of 74mA. The minimum standing current is therefore half the swing which is 37mA. A conservative value of standing current equal to 50mA is used.

The quiescent collector to emitter voltage $V_{ceq}$ can be found from:

$$V_{ceq} = V_{ce(sat)} + I_q R_L + I_q R_e$$

where $V_{ce(sat)}$ = Bottoming voltage
$R_L$ = Load resistance = 17Ω
$R_e$ = Undecoupled emitter resistance = 27Ω
$I_q$ = Quiescent collector current = 50mA
= Peak signal current

A measured value for $V_{ce(sat)}$ at 30Mc/s is approximately 3V when $V_{ceq} = 3 + 0{\cdot}05 \times 17 + 0{\cdot}05 \times 27$

$$= 6{\cdot}35\text{V}$$

The collector dissipation is:

$$P_c = I_q V_{ceq}$$
$$= 0{\cdot}05 \times 6{\cdot}35$$
$$= 0{\cdot}32\text{W}$$

The AFY19 transistor fitted with a Redpoint 5F TO-5 heat sink has a thermal conductivity

$$\theta_{ja} = 0{\cdot}1°\text{C/mW}$$

With a maximum junction temperature of 90°C and a maximum ambient of 33°C, the maximum allowable dissipation is:

$$P_{c(max)} = \frac{90 - 33}{0{\cdot}1} \text{ mW}$$
$$= 0{\cdot}57\text{W}$$

The dissipation is therefore safely below the maximum value. The circuit diagram of the driver stage is shown in Fig. 6. The supply voltage of 20V has to be reduced for the driver stage and this is done by a resistor in series with the emitter, which produces a design of high thermal stability. The base bias is supplied by a divider chain as the base current of this stage is only 1mA.

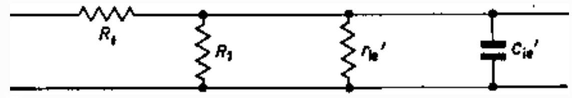



Fig. 7. Equivalent of feedback circuit

### First Stage

The power requirements of this stage are small. The drive voltage required by the driver stage is equal to its output of 1·25V peak-to-peak divided by its gain which is ×6. This is equal to 0·25V peak-to-peak. The input resistance of the driver stage is 25Ω. The current swing required is therefore:

$$I_S = 0{\cdot}25/25$$
$$= 10\text{mA peak-to-peak}$$

The transistor used is the AFY19, as used for the driver stage.

Without a heat sink the thermal conductivity is:

$$\theta_{ja} = 0{\cdot}25°\text{C/mW}$$

With a maximum junction temperature of 90°C and a maximum ambient of 33°C the maximum dissipation is:

$$P_{c(max)} = \frac{90 - 33}{0{\cdot}25} = 228\text{mW}$$

Generous operating conditions are used:

$$I_{cq} = 15\text{mA}$$
$$V_{ceq} = 10\text{V}$$
$$\text{when } P_c = \frac{10 \times 15}{10^3} = 150\text{mW}$$

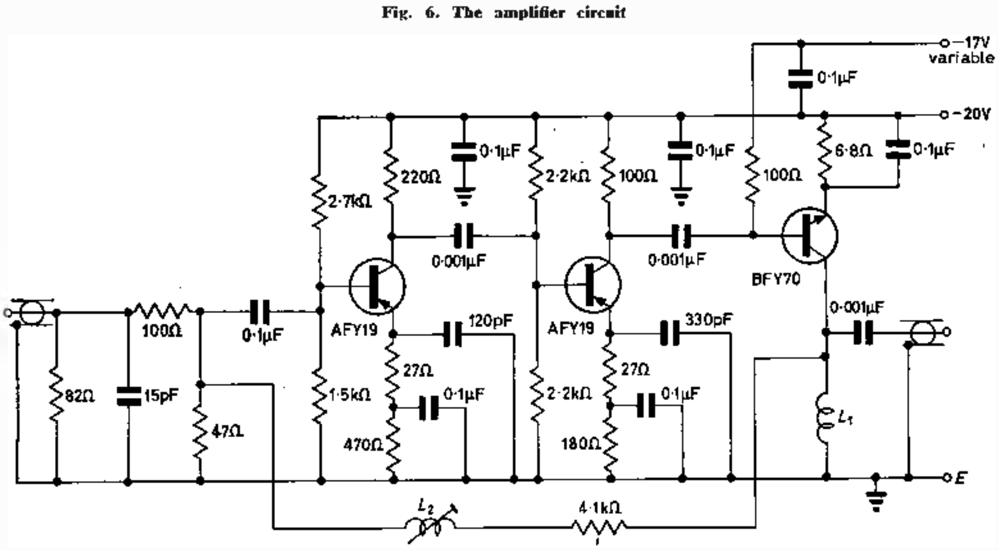
### A.C. Design

The greatest danger of instability lies at the low frequency end of the band where the loop gain is highest and the amount of feedback is greatest. Since a very low cut-off frequency is not required the simple expedient of the dominant lead is used where the time-constant of one lead is made much smaller than any of the others. This ensures a near 6dB/octave fall-off which is inherently stable.

The high frequency design can be considered in two parts. The input and driver stages with their individual series feedback loops and the output stage with loop feedback. The value of the feedback resistors in the input and driver stages can only be calculated if the source and load impedance of each stage are known, together with the parameters of the transistors. The accuracy of such data is sufficient to determine the feedback resistors approximately, the actual value is best found empirically together with the values of peaking capacitors in shunt with the feedback resistors.

If the first two stages are considered to have constant gain over the band the amplifier with loop feedback is an approximation to a two lag feedback system, the output transistor contributing a single lag. The

Fig. 6. The amplifier circuit



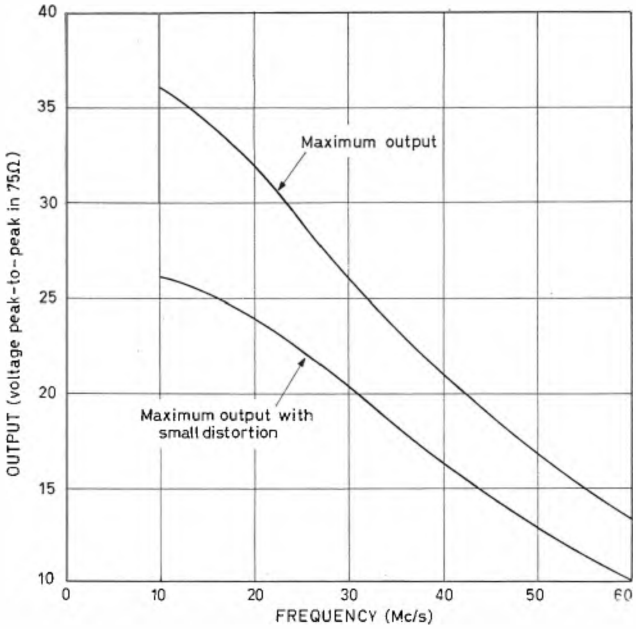



**Fig. 8. Output voltage**

second lag is formed by the feedback resistor and the input impedance of the first stage which consists of a resistor and capacitor in parallel. A two lag feedback system can exhibit a frequency response ranging from an aperiodic delta function response to a single high $Q$ tuned circuit response depending on the ratio of the two lag time-constants and the loop gain. In the amplifier the two lags are approximately 100nsec due to the output transistor and 2·4nsec at the input to the first stage. The first lag is:

$$T_1 = 1/f_{bfe}$$

where $f_{bfe} \simeq 10$Mc/s (measured)

i.e. $T_1 = 100$nsec

The second lag $T_2$ may be found from the equivalent circuit of Fig. 7. $R_f$ is the feedback resistor of 4·1kΩ. $g_{ie}'$ and $C_{ie}'$ are the input conductance and capacitance as modified by the impedance feedback in the first stage. The resistor $R_1$ is the shunt resistance formed by the input attenuator equal to 35Ω.

The input admittance of the first stage is equal to:

$$g_{ie}' \ll 1/R_f$$
$$C_{ie}' = 68\text{pF}$$

The time-constant $T_2$ is then:

$$T_2 \simeq C_{ie}' R_1 = 68 \times 10^{-12} \times 35 = 2{\cdot}4\text{nsec}$$

The gain without feedback is found by measurement to by ×300 and with feedback is required to be ×70 neglecting the input attenuator.

when since $G_{feedback} = \dfrac{G}{1 + G\beta}$

$$\beta = 1/90$$

The value of $R_f$ is found from the relation:

$$\beta \simeq R_1/R_f, \text{ since } R_f \gg R_1$$

when $R_f = 3{\cdot}15$kΩ.

In practice a value for $R_f$ of 4·1kΩ was found to give the desired result.

The loop gain $G_L$ is the product of the gain without feedback $G$ and $\beta$ which is:

$$G_L = G\beta = 300/90 = 3{\cdot}3$$

Thomason[1] gives the critical condition for an aperiodic delta function response as:

$$T_1/T_2 = 4G_L, \text{ if } G_L \gg 1$$

The minimum value of $T_1/T_2$ in this case is therefore 13·2. The value used is 42 which is greater than the minimum. The above calculations are, however, approximate as the amplifier only approximates to a two lag system and the loop gain $G_L$ is not very large compared to unity.

**Performance**

| | |
|---|---|
| H.T. Supply: | −20V, 270mA |
| Bias Supply: | Variable 3V supply adjusted to produce correct collector current in $VT_3$ |
| Gain: | 27dB at 30Mc/s |
| Bandwidth: | −3dB, 1·5Mc/s to 60Mc/s<br>−0·5dB, 3Mc/s to 50Mc/s |
| Load: | 75Ω |
| Output Voltage: | See Fig. 8 |
| Input Impedance: | 50Ω ± 2Ω |
| Physical Size: | 11 × 8 × 3cm |


**Acknowledgments**

The work which resulted in this note was carried out on a Ministry of Aviation Contract sponsored by R.A.E. Farnborough.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## Automatic Diode Assembly

A group of automatic assembly machines for the manufacture of miniature glass-encapsulated silicon/germanium diodes at the rate of 2 000 per hour has recently been completed by the Bader Machinery Co. Ltd of Weybridge, Surrey.

The group consists of a duplex beading machine supplying the necessary components for the preparation of both the whisker and the body sub-assemblies and two separate assembly machine groups which process these sub-assemblies.

One of these groups, used for the preparation of body sub-assemblies, consists of a body sealing machine, body washing, a crystal and solder feed machine and soldering ovens.

The other group deals with the whisker sub-assemblies and consists of a deborating machine, a whisker welding and forming machine and a whisker pointing machine.

The separate sub-assemblies are brought together for hermetical sealing by a machine when the whiskers are compressed to a high degree of uniformity and are pulsed during the sealing process.

The assembled diodes are then passed on to a further machine which carries out preliminary electrical tests and finally to a tinning machine and classifying machine which completes the assembly process.

Throughout the whole process, the components in their various stages of assembly are mounted on specially designed 50-position heat and acid resistant magazines. These magazines are provided with indexing points for accurate location on each machine and can be used in either the upright or inverted position, ensuring that the components are not touched by hand during assembly.

This group of machines allows a wide range of this type of diode to be produced using soft or hard glass, beaded or slugged lead-wires of various diameters and lengths, together with varying whisker shapes and metals.

# An Engineered Package for Microminiature Elements

By A. J. Mayhew*, B.A.

*After a critical review of prevailing practice in digital engineering, the features required of an ideal module are stated and a new package is presented to meet them. It is shown that high packing densities can be achieved with discrete wiring and circular-shaped components and that compatible modules can employ multi-layer circuits. Several varieties, old and new, of the latter are briefly discussed as is the application of automation to their design, construction and test. The development potential of the package is shown to be considerable.*

(Voir page 62 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 69)

THE pursuit of increased reliability and reduced cost led, a few years ago, to the introduction of integrated circuits—complete logic elements in the same shape and size package as formerly housed only a single transistor. Managements with large investments in plant for making ordinary printed wiring boards, and a distressing conservatism on the part of development engineers, have led to the present position where the design of components has completely outstripped the design of methods for mounting and interconnecting them.

Today, throughout this country and most others, digital equipment is being made almost entirely of standard printed boards, single or double-sided, with edge-connector sockets which are interconnected via the traditional cable harness. Occasionally one finds smaller boards fixed to the larger ones at right-angles, showing that plain board with only a two-dimensional circuit is too restrictive and that the designer needed to use the third dimension. One also finds some boards with little or no printed circuit but with components connected to terminals staked in the board interconnected by discrete wires, showing the unsuitability of printed circuits for short production runs or the rapid construction of prototypes or 'specials'.

Other criticisms of the genre are of the soldering process still almost universally employed and of the problem of heat removal. Despite its long history, the soldered joint is still the least reliable of the available alternatives and its use with eyelets for through-the-board connexions is even worse. To keep the heat, necessarily involved in making it, from cracking glass seals, components have to be stood off the board on insulated spacers, which helps neither mechanical soundness nor close-packing. Attempts are being made to use welded joints with boards clad with other metals than copper, but the repairability of such boards is doubtful and the packing density achievable is probably lower.

Heat removal, almost always an afterthought, is always by blowing or sucking air through the equipment. Space for air to pass through is space lost to equipment packing, and the tendency is to pack tightly and then blast air through with large fans. Apart from problems of dust and vibration, the fans can easily consume as much power as the equipment and themselves take up space.

The method by which the boards are now made necessitates a long, complicated, production line which cannot readily brook the disruptions caused by sudden urgent demands for small numbers of new types of board for development. In any case, the conventional board, even double-sided, cannot accommodate the number of conducting paths for the microcircuits that could otherwise be reasonably carried by the board.

**Design Aim**

There is therefore a need for a new engineering form designed from the beginning to overcome the problems posed by integrated circuits and flexible enough to take advantage of any new techniques that may be developed. The aims of the design may be detailed as follows:

(1) *Heat Removal*

Efficient and convenient means must be provided to remove the heat dissipated by the elements and/or control their working temperature. This is more important than ever with the new components because of the greater concentration of heat due to the high packing densities achievable.

(2) *Mechanical*

The elements must be soundly supported and the basic module as a whole must be robust.

(3) *Maintainability*

The individual elements must be separately removable for repair of the module. This is important since, if it cannot be repaired, too large a module would be too costly to throw away. Conversely, if the module is easy to repair and only the faulty parts are discarded then it can be made large, which improves packing density.

(4) *Packing Density*

The number of elements per cubic foot should be as high as possible consistent with meeting the other requirements.

(5) *Ease of Design*

For the maximum economy in prototype work and short production runs, the module must be easy to design and specify. Close-packed equipment calls for greater care in design in any case, since there is less free space to accommodate modifications.

(6) *Reliability*

The module wiring should employ reliable jointing methods, and the minimum number of joints of all kinds.

(7) *Flexibility*

The module should be economic to make in short or long production runs, with discrete components or integrated circuits and with discrete or preformed wiring.

(8) *Development Potential*

It should be possible, as they develop, to use new

* *The Marconi Co. Ltd.*

techniques and components to improve the module's performance in any or all of the above respects.

(9) *Screening*

It is an advantage for the module to be shielded from interference by a metallic screen.

(10) *Form of Element*

Engineering becomes very easy when all, or most, components are the same shape and size. This is not essential but is of value; as an example, integrated circuits, transistors, potentiometers, coils, transformers and relays may be had in TO-5 cans or within the same outline. The current design of flat package is certainly not a good one with which to try to meet the design aim.

fitting to the chassis which is already loaded with elements. The sound mechanical mounting of the latter and the large holes in the tagboard or m.l.c. make it easy to fit these items over the leads of the elements. These leads are now connected to the terminals.

Connexions to and from the package may be made in several ways. Conventional multi-way plugs or sockets may be fitted by brackets to the bottom of the chassis and connected to the internal circuit by discrete wires. Edge-connector pads may be formed, or additional terminals of a suitable shape may be fixed on tagboard or m.l.c.

An overall metal screening can, finished matt black, is securely screwed to the chassis plate.

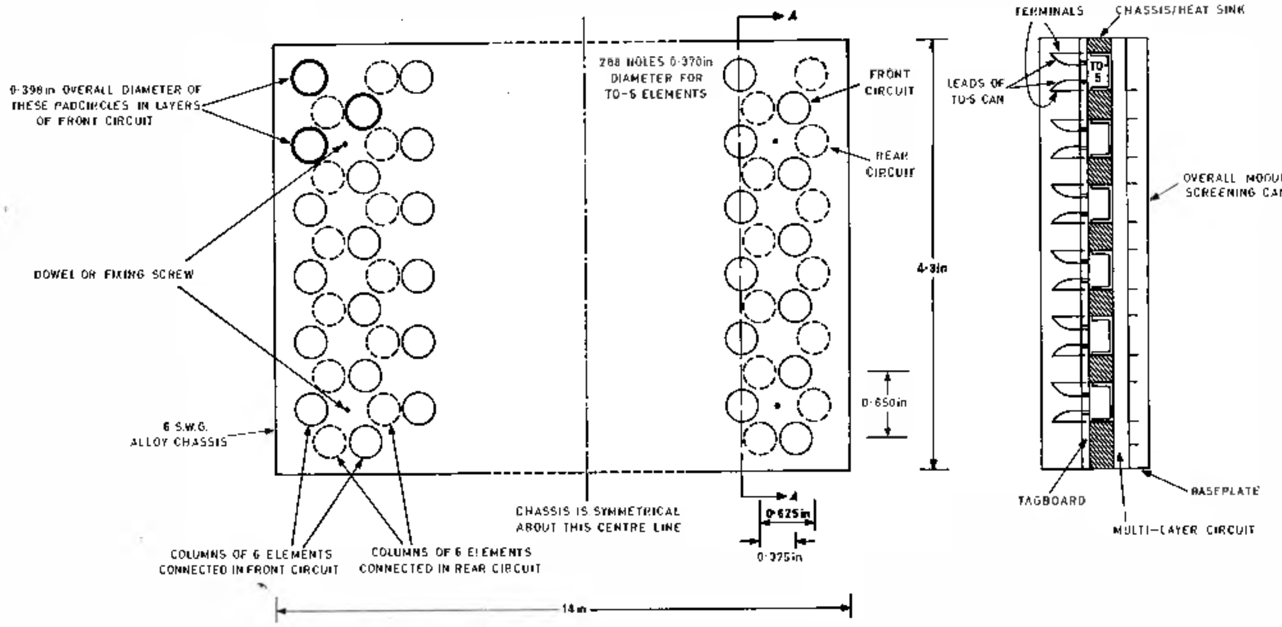



**Fig. 1. Package for TO-5 type elements, plan and section**

### The New Package

The first example, Fig. 1, shows a package for TO-5 type elements. These are inserted in a metal plate which serves as a common heat sink and the chassis of the module. The plate is approximately 0·200in thick and may be a casting or punched from sheet. To achieve mechanical support for the cans despite their $\pm 0{\cdot}015$in diametral tolerance, and to assist heat transfer, spring collets are fitted to them. The TO-5 cans are inserted from either side of the chassis in alternate positions. The cans on the two sides of the module lie on two interleaved systems of equilateral triangles. It can be shown that if the holes in the chassis just touch then the minimum mounting pitch of the elements on each side is $1{\cdot}5d \times 1{\cdot}732d$, where $d$ is the hole diameter. The pitch actually achievable depends on the minimum wall thickness desirable between the holes, and this will depend on the method by which the chassis is made and the tolerances involved.

The sectional view, Fig. 1, shows the construction of the module. The leads from the elements in the chassis protrude through large holes in tagboards or multi-layer circuits. Surrounding each such hole, groups of 8, 10 or 12 terminals are fixed to the tagboard or multi-layer circuit (hereafter abbreviated to m.l.c.). In the tagboard case, for prototypes or specials, the terminals are interconnected with discrete wires before fitting to the chassis. Thus, in both cases, the wiring between terminals is complete before

It is well known that a large basic module helps to increase packing density. Since the package described here has such a strong 'backbone', there is very little limit to its size. This means fewer identical modules in a given equipment but does not, in this case, mean that standard circuits cannot be employed. On the one large chassis may be mounted any number of part-circuits, each of which may be a standard in wide use, such as a 6-bit register. These need not all be of the same form, for instance there might be in one module some standard register units with m.l.c.'s, and decoder units with tagboards and discrete wires and even with discrete components. Between standard and/or non-standard sub-units within the module, through connexions can be made by twin-wrapped links or soldered, wrapped or welded discrete wires or other means. It can be seen from Fig. 1 that apart from the two can-position grids with equilateral spacing, there is in this pattern a third such grid of positions which can be used for dowels or fixing screws.

### Production Considerations

The maximum flange diameter in the TO-5 specification is 0·370in. The mounting pitch of the cans on each side of the module of Fig. 1 is shown as $0{\cdot}650 \times 0{\cdot}562$in which means that the holes in the chassis are at 0·375in centres. The module height allows for 6 cans in each column so that each 0·562in of length mounts 12 cans, and a 14in module, as suggested, would therefore house 288 cans.

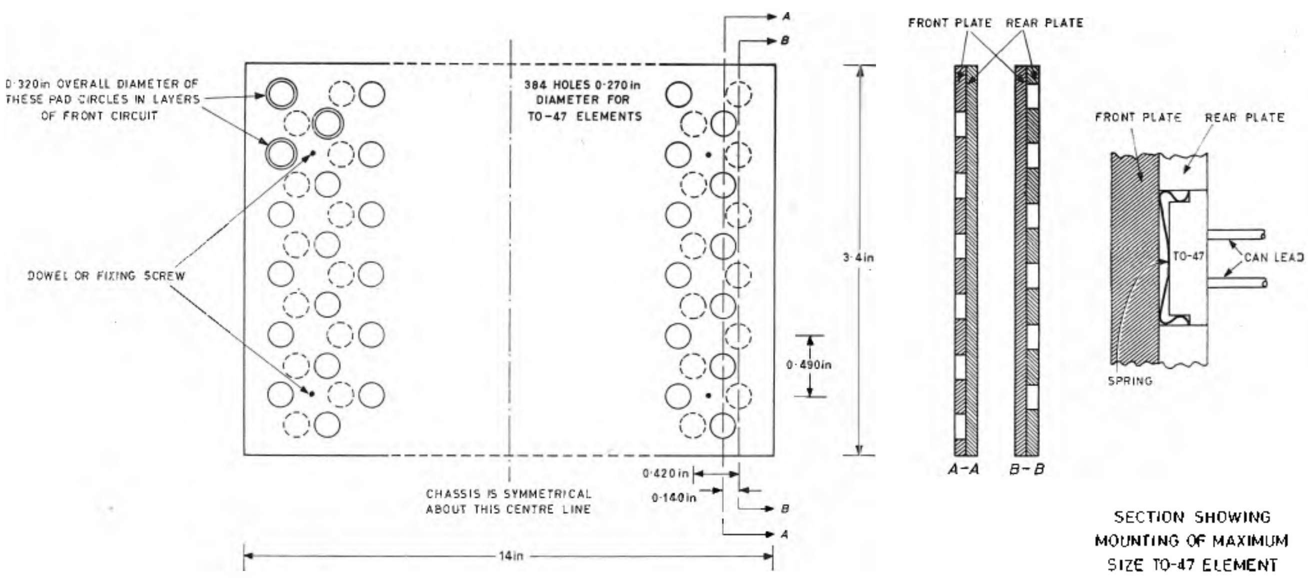



**Fig. 2. Package for TO-47 type elements, plan and section**

With the correct symmetry, the can position grids of the two sides are identical.

To accommodate the locating tag which projects from the flange of the TO-5 can, the punch which produces the 0·370in diameter hole must also produce an added slot in one direction 0·045in long and 0·034in wide. These can be punched out using a template on a Wiedeman press. Because of the identical grids of holes for the two sides of the module, and the symmetry, the template need have only one grid of holes in it. When these have been punched, the template is turned over and the other grid is punched.

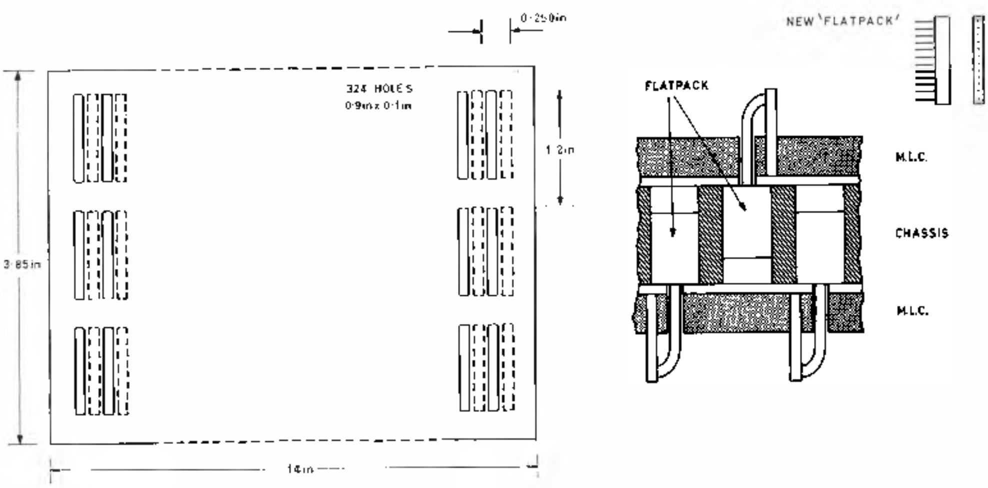
The same template can be used to punch the holes in the tagboards or m.l.c.'s. The leads of the TO-5 header are 0·018in maximum diameter on a 0·200in p.c.d. giving a 0·218in overall diameter. The holes in the tagboards or m.l.c.'s are therefore made 0·228in diameter. The terminals would be staked on a 0·328in p.c.d. Twelve pins may be required on this pitch circle, in which case they will be at 0·086in centres. A suitable size of pad in the m.l.c. layers would be 0·070in diameter, leaving 0·016 between pads and giving an overall diameter for the pad circle of 0·398in.

As suggested above, the elements on one side of the module are 0·650in apart. The space between adjacent pad circles is therefore 0·252in. With 0·020in tracks and 0·010in spaces, one can therefore fit up to 8 tracks between them.

If the TO-47 types of can be used, a further economy becomes possible in that the chassis may now be made as two identical plates, see Fig. 2, one of which is reversed before they are joined together. Support and thermal contact between can and chassis can now be established by a compound spring as shown. The limitation to 8 pins on this header permits a reduction in the p.c.d. of the pads in the m.l.c. to 0·250in, which, together with the smaller hole in the chassis, permit the reduction of the can mounting pitch to 0·485in × 0·420in. There is therefore 0·165in between adjacent pad circles for conductors. A 14in module as suggested would house 384 elements.

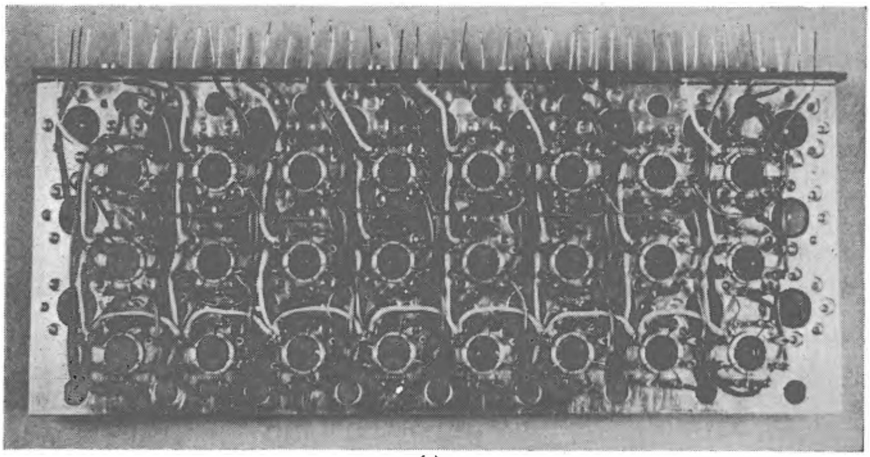
With punched holes in the chassis, other shapes than circles are feasible and a proposed new 'flatpack' with 12 connexions brought out in line on 0·075in centres can be accommodated as shown in Fig. 3. A 14in module could house 324 such elements in 54 columns of 3 on each side of the chassis. With leads on this spacing it is obvious that pad sizes must be reduced at least to 0·050in or possibly to 0·030in diameter. The latter would leave 0·045in for tracks to pass between them. The leads from the package would project through a line of 0·030in diameter holes in the m.l.c. or tagboard as the case may be.

**Fig. 3. Package for new flatpacks, plan and section**

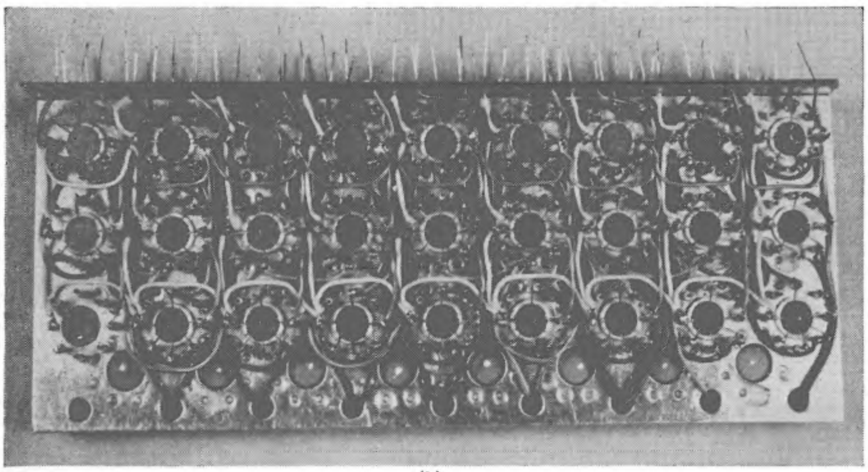



### A Prototype Package

In the early stages of the development of this packaging scheme it was decided to make a prototype to demonstrate what was achievable with the techniques that were already to hand. This is shown in Fig. 4. The



(a)



(b)

**Fig. 4. An early prototype**
(a) *Rear view* (b) *Front view*

chassis is approximately 7in × 3in and can accommodate 51 TO-5 cans in a rather simpler packing geometry than that referred to earlier.

The terminals are of 23 s.w.g. p.v.c.-covered wire cut and stripped to length and cemented with Araldite in holes drilled in the chassis. They are on a larger p.c.d. than is the case with later designs so that TO-5 elements can be inserted and withdrawn from the chassis through the rings of 8 terminals. No collets were available to take up the tolerance on the can diameter so the holes in the chassis were drilled to a 0·0005in tolerance to make the cans a good push fit. Two hundred TO-5 cans, both logic elements and transistors, from a number of manufacturers, were measured and all found to be within such a dimensional tolerance.

This module is a working model of a 13-bit Gray-Binary transcoder and register and uses 50 Fairchild Micrologic DCTL elements. The terminals are interconnected with ordinary 28 s.w.g. p.v.c.-covered wire in 5 layers with a different coloured wire for each layer for reasons given below. This work was done by an engineer with no special tools and the joints are soldered.

As the method of connecting to and from the module was not an issue at this time, the input and output leads are merely the ends of the 28 s.w.g. wires passed through holes in an s.r.b.p. baseplate. They are therefore rather fragile but would ultimately be proper terminals in an insulant or metal baseplate.

This prototype module, after allowing space for intermodule wiring, permits overall equipment packing densities of about 2 700 logic elements per cubic foot. It demonstrates clearly how easy engineering becomes when all components are the same convenient shape and size. It also shows that with this order of packing density there is plenty of room for discrete wires for interconnexion. Because of a mistake, one side of the module when complete was found to be wrongly wired. It was completely stripped and re-wired, thus proving the ease with which this type of module may be modified and hence its suitability for prototype work.

The total cost of this example, hand-made throughout, was 30 man-hours plus the cost of components and materials. It is the equivalent of 14 standard 6in × 5in printed boards using discrete components, together with their edge-connector sockets and cable harness. All components are soundly mounted yet can easily be removed and replaced. Fault tracing is facilitated by the ability to disconnect any lead of any element and to re-connect it. It is worthy of note that the worst-case dissipation of the 50 Fairchild elements is just over 2W and they use a single power supply.

**Multi-Layer Circuits**

The merits of pre-formed wiring require no advocacy to engineers still wedded to ordinary printed boards but the latter, as is explained above, are inadequate for the interconnexion of integrated circuits. The module under discussion, with its absolutely standard fairly coarse grid of terminal positions, is obviously well suited to employ m.l.c.'s of any of the types already suggested. These, and some new ones, will now be discussed briefly.

Firstly one must point out a mistaken impression that seems to have influenced all the designs of m.l.c.'s to date. This is that one must be able to connect between layers. The wiring of any device may be specified in the form of 'Clusters', a cluster being a group of points which must all be connected together but which do not connect to any other points. This form is very suitable for the design of m.l.c.'s and it follows from the above definition that *if each cluster is completely realized in one layer, then connexions between layers are not necessary.*

This point is illustrated in the prototype module of the previous section shown in Fig. 4. This was wired in layers, using different coloured wire for each layer. The first layer, in brown, contains a number of clusters whose wires are so laid out that no wire in one cluster crosses a wire of any other cluster. (Thus the layer could be printed in this pattern). The second, red, layer contains clusters whose wires cross those in the brown layer, but again no wire of one cluster crosses a wire of any other cluster in this same layer. The wiring of the third, fourth and fifth layers, in orange, yellow and green, satisfies the same criterion. It can be seen from the module that no terminal has more than one colour of wire connected to it: i.e. *no terminal connects to more than one layer.* There are up to three connexions to a terminal but they are all in the same layer and when this is translated to a printed layer they become one track to be connected to the terminal.

Thus the criterion for the design of an m.l.c. may be stated as: So long as the number of layers is not too strictly limited, no termination need connect to more than one layer. By this, the problem of multi-layering is changed from one of connexion between layers to one of establishing connexion between a terminal and any one of the layers, which opens up many new lines of investigation. Some new designs will be described after a brief review of the earlier ones.

(1) Clearance Holes

The layers are produced separately as single-sided printed circuits on thin laminate and, as usual, holes are punched in the connector pads for the component leads. Also larger or clearance holes, almost equal to the pad diameter, are punched in all layers above the first at all positions where connexions are to be made to pads in lower layers. Thus when all the layers are stacked in a jig and registered, each connecting pad is accessible through the clearance holes in the layers above it and soldered joints may be made between pads and component leads. The number of layers is usually limited to about 6 so that its usefulness is limited, but this technique is the basis of a number of more powerful designs.

(2) Eyelets and Rivets

Again separate layers are produced and stacked and holes are drilled at all terminal positions. Eyelets or rivets are now inserted in all holes and connexion between them and the exposed rings of tracks in the holes is established by either expanding the eyelets or rivets or by heating them to melt a solder coating. Again the number of layers is limited and the joint made is not reliable enough and impossible to inspect.

(3) Plated-through Holes

A number of ways of using this technique are possible but the following is fairly typical. Once again, layers are stacked after etching and stuck together. Holes are drilled through at all pad positions and then an electroless process is used to deposit a very thin layer of copper on the surfaces in the holes (and on the outside laminate surfaces). This layer is built up to the desired thickness electrolytically and then the required wiring patterns are etched on the outside laminate surfaces. The process, if reliable, is expensive and the number of layers is restricted to about 6.

(4) The Intellux Process

This is an American patent process for making m.l.c.'s electrolytically. It consists of building up alternate layers containing terminal pads only and terminal pads plus their interconnecting tracks, starting with a metal baseplate which is finally etched away. When a layer has been deposited, the spaces between pads or tracks are filled flush with insulating resin and cured and the process repeated for the next layer. A typical layer is 0·006in thick, of which 0·0015in is the interconnecting tracks, and takes about a day to lay down, the major time being in the plating baths and curing ovens. Up to 6 layers are possible and the process is obviously rather expensive.

(5) Welded Matrices

This technique uses a number of welded wire matrices to form the required wiring pattern. A single matrix consists of two layers of conductors running at right-angles to one another, separated by an insulating sheet of Mylar. The latter is prepared from scaled drawings reduced to final size photographically. On it are depicted the wiring paths required and the positions at which holes must be punched for welded connexions between perpendicular conductors. It also carries instructions as to the type and size of conductor wire or ribbon to be used and it serves as a part of the holding fixture during fabrication and becomes an integral part of the final assembly. As each layer is added to the stack, its ends are welded to wire-wrap posts or welding terminals. Finally, the entire stack is encapsulated.

From the above techniques a number of other ideas have been suggested, some of which are the subjects of patent applications.

(6) Diffused-joint Terminals

An m.l.c. is prepared by the plated-hole technique. Slightly over-size pins in the form of split bushings with either a weldable or wrappable tab are then inserted in the holes. Prior to insertion, the portion of the pin coming in contact with the hole is coated with indium. The whole assembly is now heated to initiate the indium diffusion process which will continue at room temperatures. The indium diffuses into the plating in the hole and into the terminal, forming a new alloy which joins and is actually an extension of both the plated hole and the terminal. Although the alloy is formed at a relatively low temperature, its melting point after diffusion is about 950°C and its shear strength about 28 000 lb/in$^2$.

(7) Welded Tubelets

The separate layers are prepared as for the clearance hole technique. On each layer, to every connexion pad present, a small thin-walled metal tube is resistance-welded. The layers are then stacked and stuck together. Component leads may be passed through the tubelets or through adjacent holes and connected to the former by soldering or welding.

(8) Welded Terminals

This is similar to the tubelet system but in this case solid terminals are welded to the connector pads, using a pulse/arc welder. In pulse-arc welding the work pieces to be joined are temporarily positioned together and collectively serve as one electrode of the welding circuit. The other electrode, which is fixed and does not touch the work pieces, is of a high melting point tungsten alloy. When the power supply is triggered a burst of r.f. energy ionizes the air gap, then the output transformer delivers the energy stored in the capacitor banks, heating the work pieces to the molten state. When the weld pulse ends, the melt cools, leaving a welded joint between work pieces.

(9) Welded Split Terminals

This differs from the last system in that parallel-gap resistance welding may be employed. The terminals are of a sandwich construction with a very thin layer of insulation down the middle. The electrically isolated metal sides are gripped by the jaws of the welding head so that the lower ends of the halves of the terminals become the effective welding electrodes. The terminal is pressed against the layer connecting pad and the power supply triggered. When the weld has cooled, the jaws are released, leaving the terminal fixed to the pad.

The layers may then be stacked or it may prove possible to stack them first and then attach the terminals. Once welded to a pad, the fact that the terminal has a layer of insulation down the middle is irrelevant to making a soldered or wire-wrapped connexion to it. Since the halves are now electrically connected a welded joint may be made, also by parallel-gap welding, between a component lead and either half.

(10) Formed Metal Tabs

In this system, conductors are fixed in place in or on sheets of plastic material which are then stacked. Connecting tabs from each layer are brought out to the top surface for connexion to component leads or to tabs from other layers. Obviously, holes must be cut in upper layers for the tabs from lower ones. It is said not to be economic in small quantities although new methods of manufacture could overcome this.

(11) Staked Terminals

Here, layers produced by any of the usual methods are stacked and stuck together and then needle-shaped terminals are driven into the assembly at all required positions and their tips peined over. The terminal, when driven in, parts the substrate but swages some of the connector pad down with it. The joint between pad and terminal is a gas-tight pressure joint maintained by the stresses locked in the substrate, like a nail driven into wood. In the typical module the terminals would be inserted in groups for the microelements of 8, 10 or 12. This is a readily automatable process but its reliability has to be measured.

The principal parameters of the varieties of m.l.c. mentioned are summarized in Table 1.

In choosing from these and any others that may be devised, what are the important criteria? Firstly, one considers reliability. In all cases this is improved with respect to the traditional cable harness by the consistency implicit in preformed wiring. Secondly, one looks for the minimum number of joints or interfaces, and the most reliable types of such interfaces as are involved. Next, one considers the types of joints which may be used between the m.l.c. and the component leads in the module: the more choice one has in this matter the better. Last, but by no means least, one considers the flexibility of the technique, in particular the possible number of layers. At present, with double-sided printed boards, design of the wiring layout is a very skilled process or very laborious, or both, because of the need to minimize the number of through-the-board connexions. If one has enough layers at one's disposal, this design work is very simple and merely consists of applying the criterion stated at the beginning of this section. It can be easily automated as explained later.

Furthermore, a power supply line can occupy a complete layer and be almost a complete sheet of copper with tabs to connect to the standard pin at each microelement position. The power supply layer can be made of relatively thick copper and thin substrate for minimum line impedance and maximum built-in decoupling capacitance to the similarly designed earth layer.

It is noteworthy that this easement of the design of the wiring layout applies to all applications which at present use double-sided boards or single-sided boards with numbers of jumper leads, and is not confined to high-density microelectronic applications.

A form of hybrid circuit, which provides a simple but worthwhile introduction to the technology of m.l.c.'s, is one in which the power supply wiring is multi-layered, but the signal connexions made with discrete wires. Apart from helping to introduce m.l.c.'s, this also forms a useful method of wiring protoypes and short production runs, especially for modules with a standard grid of component positions and standard power supply pin connexions.

If any advocacy should still be thought necessary to justify the furious effort being applied to the development of m.l.c.'s it may be summarized as follows:

(1) The packing densities possible with integrated circuits are not achievable with existing printed board techniques.

(2) The reliability of existing interconnecting techniques has not kept pace with the improved reliability of integrated circuits.

(3) Conventional connexion techniques require far too much bulk and preclude the achievement of the reduction of equipment size and weight made possible by integrated circuits.

(4) A greater proportion of an equipment's wiring can be pre-formed.

(5) The wiring layout is easier to design and the process can be assisted by machines.

**Production Sequence**

In order to show the economics of this package, it is convenient to list the steps in the production of a module. Taking the TO-5 module of Fig. 1 as an example, then the major steps involved are:

*Chassis*

(1) Cut out metal sheet.

(2) Punch one grid of 0·370in holes, using template.

(3) Reverse template and punch second grid of holes.

(4) Punch and thread dowel/fixing holes.

(5) Drill and thread fixing holes for can and baseplate.

(6) Load with logic elements.

(7) Fit tagboard or multi-layer circuits.

**TABLE 1**

| M.L.C. TYPE | JOINTS TO COMPONENT LEADS | | | | MINIMUM JOINTS TERMINAL TO TERMINAL | TYPE OF INTERNAL JOINTS | MAXIMUM NO. OF LAYERS |
|---|---|---|---|---|---|---|---|
| | SOLDER | WAVE-SOLDER | WIRE-WRAP | WELDED | | | |
| (1) Clearance Holes | * | – | – | – | 0 | none | 6 |
| (2) Eyelets, Rivets | * | * | – | – | 2 | pressure/ solder | 6 |
| (3) Plated Holes | * | * | – | – | 4 | plated | 6 |
| (4) Intellux | * | * | – | – | ? | plated | 6 |
| (5) Welded Matrix | * | * | * | * | many | welded | no limit |
| (6) Diffused Joint | * | * | * | * | 6 | plated, diffused | 6 |
| (7) Tubelets | * | * | – | * | 2 | welded | 10 |
| (8) Terminals | * | * | * | * | 2 | welded | no limit |
| (9) Split Terminals | * | * | * | * | 2 | welded | no limit |
| (10) Formed Tabs. | * | * | – | * | 0 | none | 10 |
| (11) Staked Terminals | * | * | * | * | 2 | pressure | no limit |

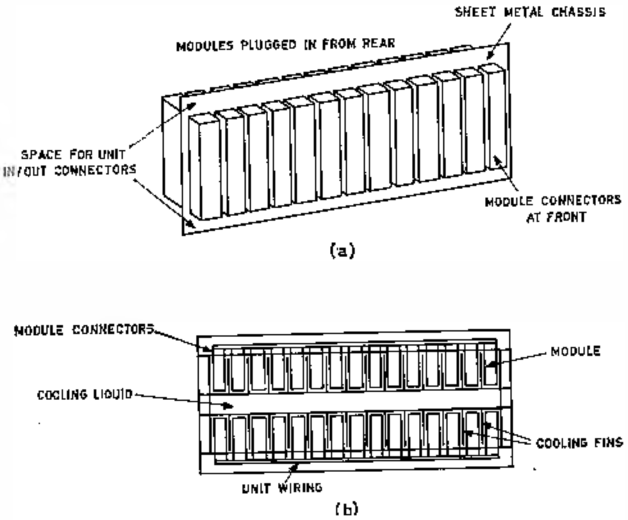



**Fig. 5. Unit and cabinet design**

(8) Connect element leads to terminals.
(9) Fit baseplate.
(10) Fit overall can.

*Tagboard*

(1) Cut out board.
(2) Punch grid of holes, using template.
(3) Stake groups of terminals, locating in grid holes.
(4) Punch insulating rear layer.
(5) Attach rear layer.
(6) Wire to schedule with soldered or welded joints.
(7) Fit to chassis.

*Multi-Layer Circuit* (One method)

(1) Print layers.
(2) Stack layers in jig and stick together.
(3) Punch grid of holes, using template.
(4) Stake rings of terminals, locating in grid holes.
(5) Punch insulating rear layer.
(6) Attach rear layer.
(7) Fit to chassis.

Some important points may be noted here. One is that the same template may be used to control the punching of the grids of holes in the chassis, tagboards and m.l.c.'s. Further, since this grid is the same in all modules, it is economic to make the template to very fine tolerances which improves the accuracy of many production processes. Another is the scope for machine aids to the design, production and testing of the modules, as discussed later.

Finally, it is important to note that the m.l.c.'s for use in this module may be made by any of the methods so far devised, as and when they become practical economic propositions. Until they do, the m.l.c. can continue to be made the old-fashioned way with discrete wires and soldered or welded joints. Moreover, there need be no alarm over a multiplicity of 'Weld Schedules', since in the discrete wire case there are only two schedules—wire to terminal and element lead to terminal—and in the m.l.c. case only the latter.

**Equipment Design**

So far the design of the module, the basic brick, has been considered, but consideration must also be given as to how they are to be used to build equipment. Fig. 5 shows one suggestion. Here a unit consists of one or more rows of modules plugged or otherwise fixed in their connectors, which are bolted to a sheet metal chassis with their long dimension vertical. Connexions to and from a unit are made via unitor plugs and sockets.

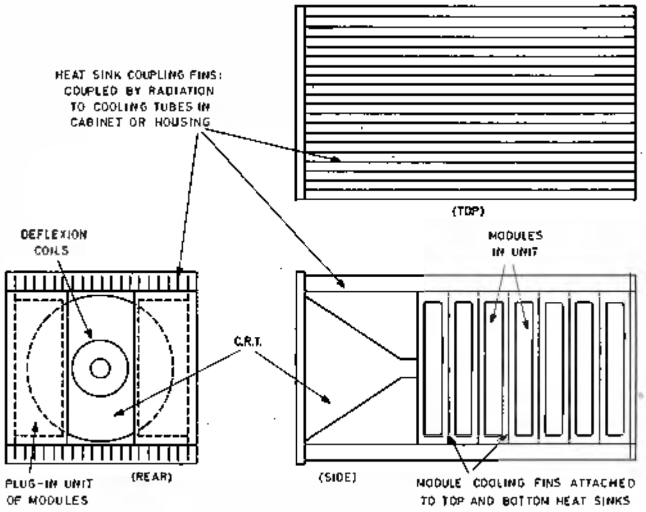



**Fig. 6. Application to c.r.t. viewing unit**

When the unit is pushed home into the cabinet, the plugs and sockets mate and the overall cans of the modules interleave with blackened metal fins which are part of a liquid cooling system fixed in the middle of the cabinet. Heat is transferred between the fins and the blackened screening cans of the modules by radiation across the small space between them. Any actual contact between fins and cans, whether due to tolerance build-up or bending of the fins, can only assist heat transfer.

In reasonable ambient temperatures the cooling liquid could be circulated by convection past a heat exchanger mounted inside or outside the cabinet. In other cases, forced circulation and complete cooling/heating systems could be used to control the module operating temperatures. If there were no adverse ambients and no great dissipation in the modules then the liquid cooling system could be omitted. An alternative method of cooling would be to mount units only in the front of the cabinet and to regard the finned structure as a large heat sink exchanging heat with air blown or sucked over the fins at the rear.

The important point is that a unit may be withdrawn from the cabinet without breaking any liquid line.

Table 2 shows the possible varieties of internal design of the module which could all have

**TABLE 2**

| CHASSIS | ELEMENTS | INTERCONNEXIONS | JOINTS |
|---|---|---|---|
| Punched with standard grid of holes | TO-5, TO-47 etc. integrated circuits or transistors + discrete components, new flatpacks. | Discrete wires | Soldered Welded |
| | | Multi-Layer circuits | Soldered Wave Soldered Bound Welded |
| Plain (Specially Drilled) | e.g. power transistors bolted to one side | Discrete wires | Soldered Welded |
| | | Multi-Layer circuits | Soldered Welded |

**TABLE 3**

| ELEMENT | MODULE PITCH (IN.) | ELEMENTS/ MODULE | DENSITY |
|---|---|---|---|
| TO-5 | $15\times 5\times 1$ | 288 | 6640 |
| TO-47 | $15\times 4\frac{1}{2}\times 1$ | 384 | 9830 |
| New flatpack | $15\times 4\frac{1}{3}\times 1$ | 324 | 8300 |

identical outward appearance and standard dimensions and mounting arrangements.

Taking account of the possible mixtures of these internal designs within a single module, this is obviously a very accommodating package and, as suggested earlier for prototypes of high speed digital circuits the power supply distribution could be a simple m.l.c. and the logic wiring could employ discrete wires on top of it.

Two further illustrations of its application may be of interest. Fig. 6 suggests a method of constructing and cooling a c.r.t. viewing unit. The circuit is constructed in standard modules mounted as units in the usual way. The fins to which they transfer their heat are attached to large finned heat sinks covering the full area of the top and bottom surfaces of the unit. When the unit is pushed home into its cabinet or housing, the fins of the upper and lower heat sinks interleave with tubes through which cooling liquid flows. Once again the unit may be withdrawn without breaking any liquid line.

Finally, take the case of a digital azimuth take-off encoder which may require head amplifiers to distribute its outputs over long cables. These amplifiers, and possibly code-conversion circuits, could be built up in the standard form but on circular metal chassis/heat sinks to match the synchro case of the encoder, see Fig. 7.

Reverting to the standard equipment form with rows of modules with their long dimension vertical, if one inch is added to this dimension for inter-unit connexion and an inch to the height of the module for the connectors and the wiring behind them, then Table 3 shows the overall packing densities achievable with this package, expressed as the number of integrated logic elements per cubic foot (and assuming only one such element per component).

To appreciate these figures fully, one may compare them with those for a new computer using TO-5 elements and printed boards—2000—and for an earlier equipment using printed boards and discrete components—about 300 per cubic foot. Despite the high packing densities, however, every single microelement may be removed and replaced, and efficient cooling is possible.

**Automation**

Whatever the method by which the module's internal wiring is realized, the assembly will, in the long run, be cheaper if as many as possible of the design and manufacturing processes involved may be performed by machines. This is because the machines can work faster and more reliably than human operators and can incorporate built-in self-testing facilities. Examples are the wave-soldering process and the use of computers to generate wiring schedules. Automation can be applied to the design, production and testing stages of m.l.c.'s.

The designer produces a logic diagram which depicts all the logic elements and their interconnexions required to meet the specification. Each element on the diagram may be given a name and its connecting pin numbers may be added. An operator can now read the wiring information from the diagram in the form of clusters and type it on a paper tape perforating machine. From this information, a digital computer can be programmed to allocate addresses in a standard module to the names of the elements on the logic diagram so that path-lengths are minimized.

If the module is to employ tagboards and discrete wires, the computer can convert the original information into a wiring schedule for the wireman. It does so in the same way as computers are now converting clusters lists into wiring schedules for units consisting of arrays of edge-connector sockets or pinbases. The computer determines how the cluster may be realized with the minimum total length of wire, in other words it determines the start and finish points of pieces of wire. This information can be put out still in cluster groups or the computer can sort it into any preferred order for tabulation eventually for the wireman.

If m.l.c.'s are to be used, and a reasonable number of layers is possible, the computer can be programmed to allocate the clusters to layers. It does so on the criterion that each cluster is completely realized in a single layer: thus it reads the list of clusters whose wiring routes it has chosen and looks for conflicts. As many as possible are allocated to the first layer, then as many as possible of the remainder to the second layer and so on. If the number

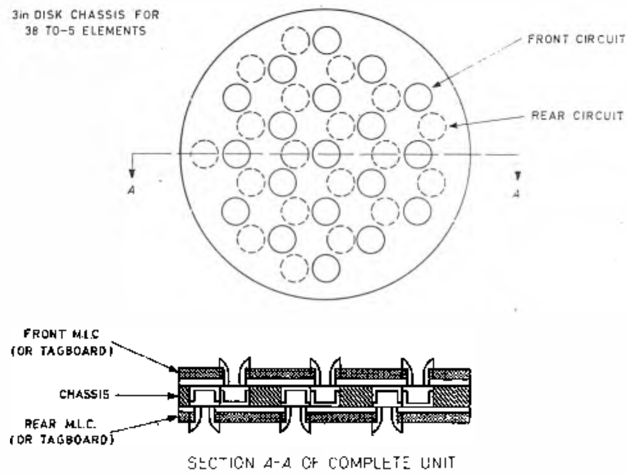



**Fig. 7. Module to match synchro case**

of layers is limited it can allocate as many clusters as possible in this way, leaving the remainder of the wiring to be added with discrete wires. The computer could output this information or, in the ultimate, it could go on to control an $X$-$Y$ co-ordinate plotting table to produce the actual artwork for the layers.

So far as production is concerned, considerable ingenuity has already been applied to the mechanization of the conventional etching process for producing printed circuits. A recent patent application describes a proposed method for high-speed mass production of both conventional circuits and of layers for m.l.c.'s and an outline of a possible automatic factory for the production of modules using m.l.c.'s with staked terminals.

Apart from the clearance hole and formed metal tab types, all the m.l.c.'s mentioned have joints or interfaces in them and all of them could be damaged in the course of manufacture. Therefore a continuity test is required. This can readily be performed by automatic machines controlled by punched tapes which can easily be prepared by the computer from the wiring information fed into it in the design phase. The tagboard with discrete wires can be dealt with in the same way.

Finally, logic tests can be applied to complete modules by automatic machines as is done now on elements and printed boards.

**Conclusions**

The properties of the ideal module have been stated as a design aim and a new package has been evolved to meet it. We may now see how well it does so.

For thermal control, efficient and convenient means have been provided which at the same time provide overall screening and mechanical soundness. Its strength and ease of repair permit a large basic module, thus improving packing density, without losing flexibility. This latter property, flexibility, the package possesses in unusual degree. It proves that high packing densities can be achieved even with old-fashioned methods of wiring, yet externally identical packages can employ preformed wiring of any type yet proposed. As a manufacturer acquires equipment and techniques, the joints may change from soldered to wave-soldered to bound (secondary wrap) to welded. The ease of design obtained when ordinary printed boards are abandoned and ways in which machines can assist in design, production and test have been shown.

TO-5, TO-47 and new flatpack elements can be accommodated which would seem to dispose of the contention that circular-packaged elements are unsuitable for computers. The reason usually given is that this shape makes it difficult to connect up without excessive lead lengths using normal printed boards. It is the latter that are wrong. The current crab-shaped flatpack is quite impossible to employ in the new package and, in any case, is far more expensive than, for example, the TO-5 encapsulation. It is difficult to seal and to wave-solder into circuit. The TO-5 is far more convenient for efficient heat sinking.

Logic elements are now available with edge speeds of a few nanoseconds. With the packing densities necessary for such fast components, printed boards must give way to multi-layer circuits of one type or another, eventually, perhaps, like striplines. Such densities will also exacerbate the heat problem, which will no longer permit of being considered as an afterthought.

The package presented in this article has sufficient development potential to be used in such future applications yet can be used economically today, with today's components and techniques, for fast or slow circuits.

**Acknowledgments**

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## West London Traffic Scheme

What is believed to be the world's most advanced fully integrated computer-controlled traffic system—developed and produced by the Plessey Automation Group—will be installed experimentally in West London under a contract awarded by the Ministry of Transport. The Plessey system will not only help to solve London's traffic problems but will also put British traffic engineering and control techniques into the world's leading position.

Under the contract the Traffic Division of the Plessey Automation Group will provide a complete range of equipment, plus the systems engineering for all aspects of road traffic management in the West London Scheme. Ministry engineers, who have prepared the specifications, will take the leading part in programming.

Although there are some larger computer-controlled wide-area schemes at present in operation throughout the world, the West London Experiment involves many entirely new features. For example, it will be the first fully-flexible scheme anywhere which combines intersection control by vehicle actuation—involving roadside detectors on each approach—together with central office computer control where overriding tactics and strategy can be imposed. Other schemes merely match this information against a series of comparatively simple pre-set programmes. In some cases, no provision is made to take into account the actual traffic conditions but the computers simply utilize programmes based entirely on historical information.

The West London traffic experiment will initially control 68 intersections in an area bounded by Hammersmith Road, Kensington High Street, Kensington Road, Knightsbridge, Sloane Street, Lower Sloane Street, Royal Hospital Road, and Chelsea Embankment. The sequence of operation is that vehicle detectors will feed information through special data transmission equipment to the central office in Westminster where there is a data scanner incorporating a computer. Data channellers route the information to the control and display units, and simultaneously to the central processor.

The central processor uses a Plessey Automation computer in the XL series, which is a synchronous, versatile, high speed machine of great reliability. The data transmission equipment uses a similar computer—and the system has been so designed that either computer can check the other. The actual processor can take over the functions of the data scanner if required, so that full display information on the situation and manual control facilities are still available.

The majority of intersections are controlled by Plessey type 54 roadside controllers, which in themselves offer very flexible control, and will operate independently should the link with the central office be broken, such as in a partial power failure. This facility eliminates the possibility of complete traffic paralysis due to a minor failure.

The data transmission outstations use integrated circuits which greatly improve reliability and also eliminate the need for routine maintenance procedures.

The equipment necessary to provide a computer controlled wide area traffic control system of the type being supplied may be broadly categorized into four major sub-systems:

(a) Central processor computer

(b) Data transmission

(c) Control and display

(c) Roadside traffic signal controllers and associated vehicle detection equipment.

The central processor and the control and display sub-systems are located at the central office, together with the central portion of the data transmission sub-system. The central office equipment is connected to each of the 68 local signal controllers via single pair GPO telephone line and a data transmission out-station unit, the latter being housed in a separate weatherproof outer case situated close to its associated intersection controller.

The 68 local signal controllers are of the Plessey type 54 or SGE type 15, both of which provide a sophisticated form of intersection control by vehicle actuation, in which traffic detectors are provided to control demands for right of way; to vary the minimum green in accordance with the number of vehicles waiting at the stop line and to control the extension of green to each approach in accordance with the speed of each vehicle and the traffic density. Additional interposing equipment has been added to make each controller compatible with the data transmission system, so that traffic data may be controlled and transmitted to the control office and control data may be transmitted from central office to the local controller.

# A Voltage Controlled Oscillator Using a Standard Integrated Circuit

By A. D. Walker*

*A simple circuit is described which will produce a relationship between input and output which is linear with either period or with frequency. The basis of the circuit is a standard integrated circuit differential amplifier. The circuit provides satisfactory operation up to 100kc/s coupled with small size, low weight, low power consumption and high reliability.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

THE simple voltage controlled oscillator may be regarded as a device which encodes a potential as a time. The relation which it produces between the input and the output may be linear with period, linear with frequency, or one of a set of non-linear functions. However, the two linear-relationships are often the most useful ones, and the circuit described here creates either of them. The principle of operation is to use a linear ramp of voltage as a timing device, and to vary the height while keeping the slope constant, or vice versa. The circuit may be regarded as a voltage to time transducer, Fig. 1; the methods by which it generates the two functions are shown in Figs. 2 and 3.

Fig. 2 shows a constant slope ramp of voltage which is stopped and reset to zero each time it reaches the value $e$. The relationship between $e$ and $t$ is the transfer function

$$t = ae \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where $a$ is a constant. The period $t$ is plainly proportional to the input voltage $e$.

Fig. 3 represents a ramp of voltage, whose slope is determined by $e$, which is stopped and reset to zero each time that it reaches a fixed value $E$. The time $t$ is now related to $e$ by an equation for the instantaneous voltage $v$:

$$dv/dt = be$$

where $b$ is a constant. This may be written:

$$v = b \int e \, . \, dt \qquad \ldots\ldots\ldots\ldots \quad (2)$$

But $v$ is only allowed to reach a value $E$, and if $e$ is held constant for this time then:

$$E = be \int dt$$

If the initial conditions are zero this yields the transfer function:

$$1/t = be/E \qquad \ldots\ldots\ldots\ldots \quad (3)$$

which shows that in this case the reciprocal of the period $t$, which is the frequency of the output, is proportional to the input voltage $e$.

The transfer functions (1) and (3) have been derived by assuming that the change in $e$ is very small during the time $t$, that is to say, the frequency of $e$ is very much lower than the output frequency of the oscillator, which is often the case in practical applications. Nevertheless it may be seen that the device of Fig. 2 is not affected by high input frequencies, in so far as a given period at the output is always proportional to the instantaneous value of the input voltage at the final moment of the period. The arrangement shown in Fig. 3 behaves somewhat differently; in equation (2), it is no longer legitimate to take $e$ outside the integral, and the ramp is stopped and reset when:

$$E/b = \int e \, . \, dt$$

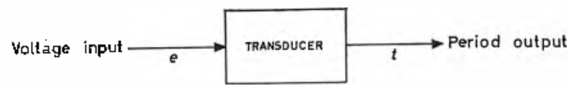



**Fig. 1. The voltage controlled oscillator as a transducer**

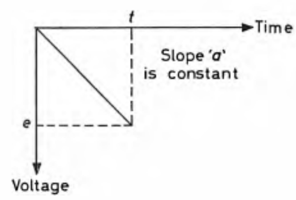



**Fig. 2. Period proportional to voltage**

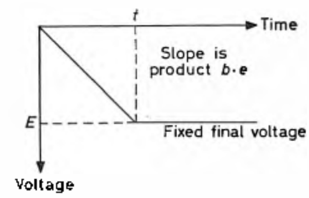
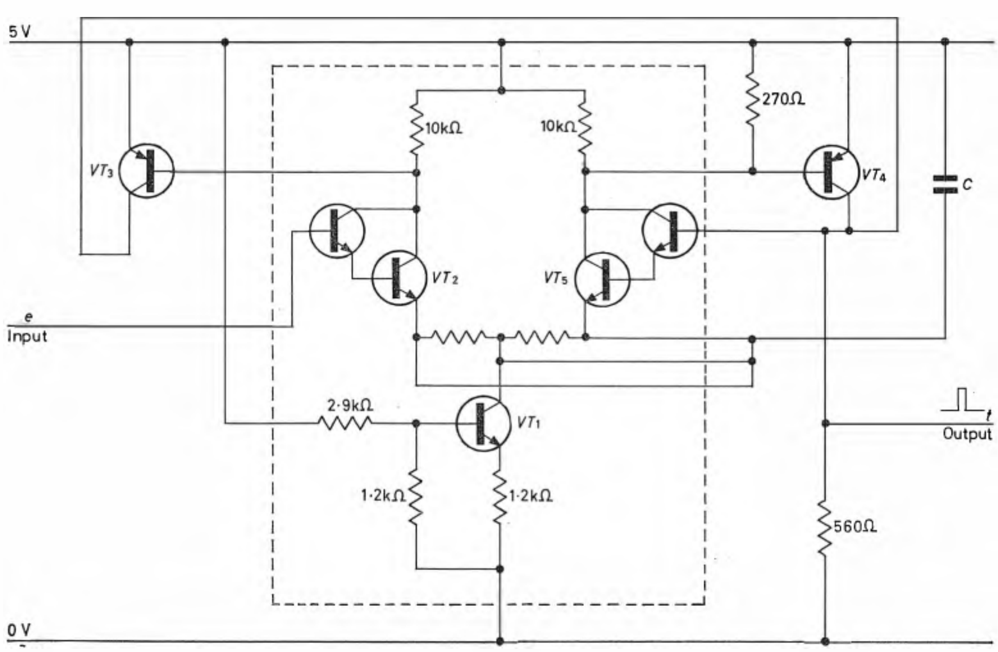


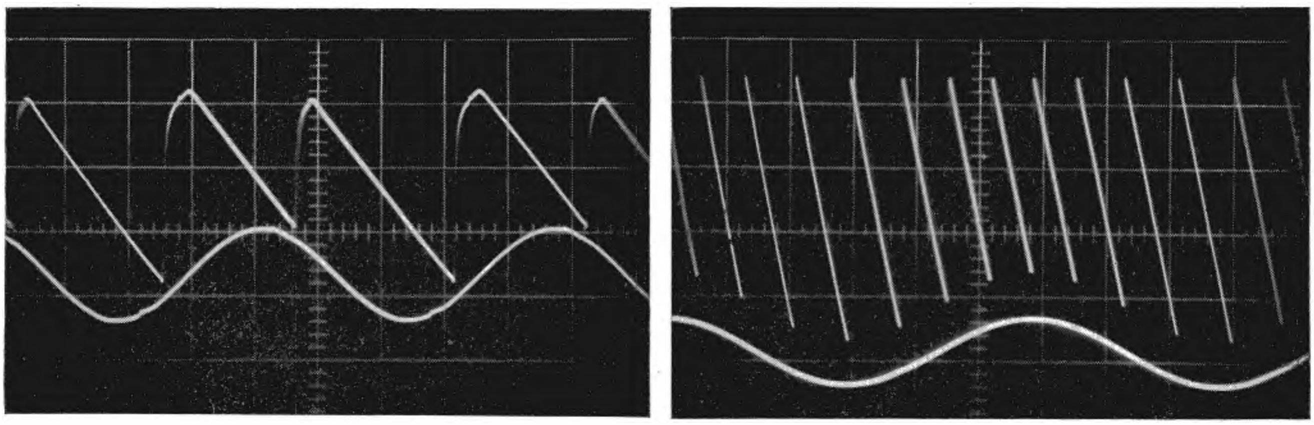

**Fig. 3. Frequency proportional to voltage**

**Fig. 4. Linear voltage to period transducer**

*The components in the dotted enclosure make up an integrated circuit*




* RCA Victor Co. Ltd, Canada.



(a) (b)

**Fig. 5. Input and capacitor voltage waveforms of the linear period oscillator**

*(a) vertical scale ½V/cm; horizontal scale 2μsec/cm* *(b) vertical scale ½V/cm; horizontal scale 50μsec/cm*

that is the output frequency is related to the area under the input voltage curve.

### Circuit Configuration and Behaviour

The circuit which is described here is designed for an application in which low power consumption, weight, and volume are important. The ramp voltage is created by extracting charge from a capacitor with a constant current source, and the charge is restored by the regenerative action of a pnp-npn transistor pair. The linear voltage oscillator is shown in Fig. 4, and its action is as follows:

If there is initially no potential across the capacitor, the current source $VT_1$ withdraws charge and the voltage $v$ drops at a constant rate. When $v$ drops below the value of the input potential $e$, transistor $VT_2$ turns on and, through $VT_3$, switches in the regenerative pair $VT_4$ and $VT_5$. These rapidly replace the charge removed from the capacitor during the ramp, but when this is completed, the only remaining current in $VT_5$ is that of the source $VT_1$; this does not provide enough voltage drop in the 270Ω resistor to keep $VT_4$ active, and the pair switches off, allowing a new ramp to begin.

The input impedance is very high, since a current of the order of nanoamps flows into the base of $VT_3$ with a very low duty cycle. The main component of the circuit is a Fairchild μC 101 integrated differential amplifier, a standard item. The output consists of a positive going pulse at each reset time. If the current source drains $i$ amperes and the capacitor is of value $C$ farads, then the transfer function (1) becomes simply:

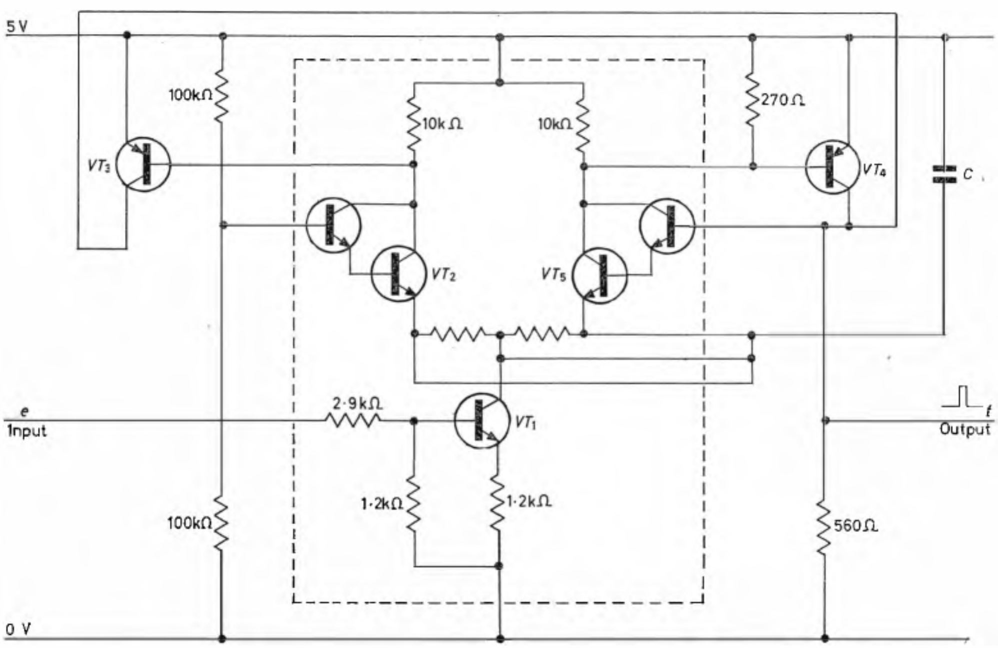
$t = Ce/i$ seconds

where $e$ is in volts. This equation connects changes in $t$ and $e$ satisfactorily, but if $e$ is measured from the upper power rail it is necessary to add a constant to the right-hand side of the expression. The lower limit on the value of $t$ comes from the speed of the resetting process; oscillograms of the ramp waveform with a sinusoidal input voltage are shown in Figs. 5(a) and 5(b), from which it may be seen that the highest useful frequency is about 250kc/s. The oscillator will function at very low frequencies. The power consumption increases slightly with frequency, and it is 12mV at 100kc/s. The weight of the components is 2·4g. The circuit operates satisfactorily over the power rail voltage range 4 to 6V, but the output period is affected by changes in this potential because of the resistor chain which determines the current in $VT_1$. The output period tends to decrease with increasing temperature, and the predominant cause is the change of the base-emitter voltage of $VT_1$.

The linear voltage frequency oscillator is shown in Fig. 6. The circuit is similar in action to the previous one except that the slope of the ramp is controlled by $e$ through the current source, while the final voltage $E$ is set up by an additional potentiometer chain at the base of $VT_2$. The

**Fig. 6. Linear voltage to frequency transducer**

*The components in the dotted enclosure make up an integrated circuit*



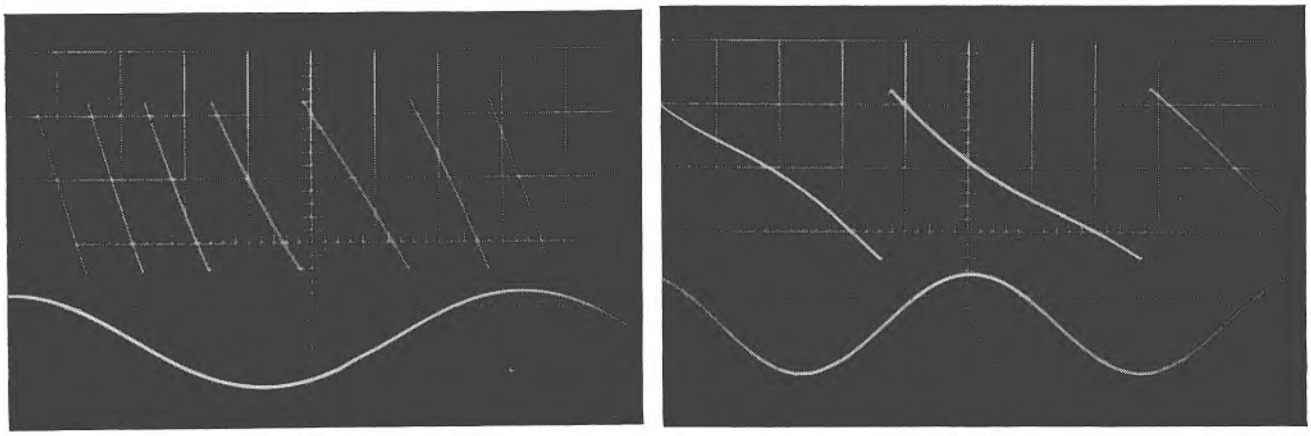

(a) (b)

**Fig. 7. Input and capacitor voltage waveforms of the linear frequency oscillator**

*Fig. 7(b) shows particularly the time integral effect which occurs when the input frequency approaches that of the output*

*(a) vertical scale 1V/cm; horizontal scale 25μsec/cm* *(b) vertical scale 1V/cm; horizontal scale 10μsec/cm*

input impedance is only of the order of 3kΩ in this case, but the output is again a positive going pulse at each reset time. If the current $i$ is a constant $k$ times the voltage $e$, then the transfer function (3) becomes:

$$1/t = ke/CE$$

This equation links the changes in $e$ and $t$, but depending on the point from which $e$ is measured, it may be necessary to add a constant to the right-hand side of the expression. This circuit is also sensitive to variation of the power rail voltage in so far as this alters the height of the ramp, and the output frequency increases slightly with temperature. Oscillograms of the voltage at the capacitor caused by a sinusoidal input to the oscillator are shown in Figs. 7(a) and 7(b); in the latter the frequency of the input waveform is of the same order as that of the ramp voltage, and the output expresses the time integral of the sinusoid during the period of the oscillator.

The linearity of both devices appears to be satisfactory if they are operated over reasonable ranges below 100kc/s, which is to say that it is impossible to detect any irregularities by measuring the input voltage with a digital voltmeter and the output period or frequency with a digital counter, both to three significant figures. However, at higher frequencies the positive edge time and the upper voltage of the ramp of the period oscillator are slightly inconsistent; this may be seen in Fig. 5(a). The frequency oscillator relies in principle on an instantaneous resetting process, so that when the positive edge time becomes an appreciable fraction of the total period, there is some loss of linearity.

### Form of the Output Signal

Although the transducers draw very little power they are unlikely to require buffer stages at the output in most applications since they produce a sharp, short pulse from a low impedance. If a square wave is required the pulses may simply be counted by a bistable circuit, and the balance of the resulting signal may be made extremely good, that is its mark-to-space ratio may be close to unity for a given input voltage to the transducer. For instance, the clock input of a Fairchild DTµl 931 integrated flip-flop may be connected directly to the output of the oscillator. The flip-flop operates from a single 5V supply and consumes approximately 30mW of power, so that the combined dissipation becomes 42mW, while the weight and volume are increased by small amounts.

### Conclusions

The basic oscillator is light, compact, consumes little power, and it is suitable for use without adaptation in many different applications, except those requiring very high frequency or tolerance to power rail variation. Within these limitations the device is linear.

While an increasing number of semiconductor manufacturers now offer to make special integrated circuits to the customer's design, this service is at present expensive unless it is spread over a large production order. In this situation the use of standard integrated circuits in unusual ways may offer some of the advantages of integration, such as reliability and compactness, at a similar cost in components and labour to that of a conventional circuit.

**Acknowledgments**

The author's thanks are due to Dr. M. A. Maclean of the Canadian Defence Research Telecommunications Establishment for his help and encouragement in this work.

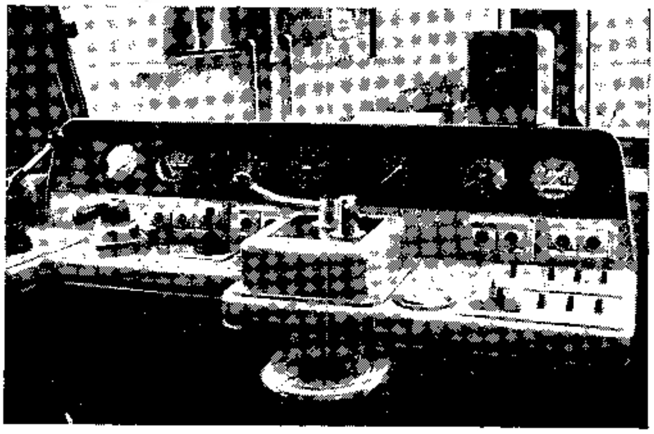
## Locomotive Speed Control

A dual range electronic speed indicator system for diesel electric locomotives has been developed by Hawker Siddeley Dynamics having ranges of 0 to 3 mile/h and 0 to 100 mile/h.

This speed indicator system contains a low speed control device which is of particular application for trains delivering coal from mines to power stations where exact speed control is required.

The driver selects the appropriate speed at which the train should run and leaves the control system to take over. The system then compares the setting with the actual speed which is measured by an inductive probe pick-up mounted close to a toothed wheel geared to the locomotive driving wheels. The electrical output of the control system is then made to control the main generator of the locomotive.

*The driver's controls of the 2750hp Brush Type 4 diesel electric locomotive. The 0 to 100 mile/h indicator is mounted in the centre and the 0 to 3 mile/h indicator above the main panel*



# A Wideband D.C. Amplifier for an Inductive Load

By M. Bronzite*, B.Sc.

*This article describes a d.c. amplifier with a linear response up to 100kc/s which is suitable for use with an inductive load. The voltage amplifier has local feedback to stabilize the open loop gain; common mode feedback to provide thermal stability; and overall feedback to linearize the response. The current amplifier also has a circuit to prevent thermal runaway. The problem of amplifier stability at high frequencies is reviewed, and a method for alleviating this problem is considered in some detail. Further discussion is given to thermal stability, and to the d.c. power supply requirements which have been met in a somewhat novel manner.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

THERE are problems associated with inductive loads which are not normally met with d.c. amplifiers. That is to say: the amplifier should have high output impedance to drive current into the load; the d.c. supplies must be relatively high to permit high voltage excursions across the load at high frequencies; at low frequencies the power handling capabilities of the output stage must be considerable (resulting from the high supply voltages); and the stability of the amplifier requires more care and attention than is normal with resistive or capacitive loads. It is felt that the amplifier to be described below has, to a large extent, overcome these problems.

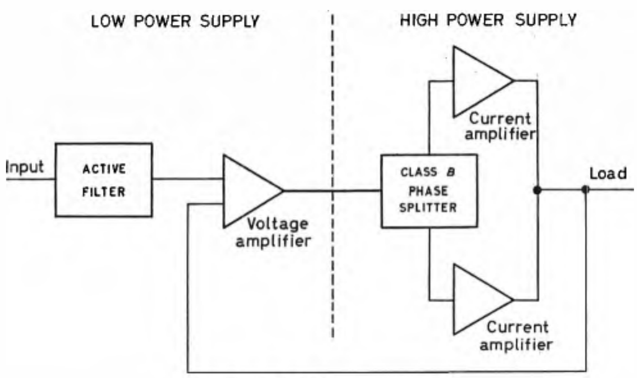



Fig. 1. The basic arrangement

The design philosophy was largely determined by the circuit specification which is outlined in Table 1. The amplifier consists basically of a voltage amplifier followed by a current amplifier with overall feedback to linearize the response. In order to reduce the power dissipation in the output transistors, it is expedient to use a class-AB output stage; which in turn requires the use of a phase splitter stage. The voltage amplifier is fed from an active filter (which is outside the feedback loop) in order to reduce the amplitude of the signal outside the operating frequency range. To facilitate the power supply requirements the current amplifier is fed from a relatively poorly stabilized high power supply, while the voltage amplifier is fed from a well stabilized and regulated low power source. The basic arrangement is shown in Fig. 1.

**Circuit Analysis**

The circuit diagram is shown in Fig. 2. Examination of the output stage shows that it is based on the structure of Fig. 3. Thus transistors $VT_{15}$, $VT_{16}$, and $VT_{17}$, with associated resistors are equivalent to a pnp power transistor. Resistors are included ($R_{49}$ and $R_{50}$) to ensure that, at low currents, $VT_{15}$ and $VT_{16}$ are not working in a starved condition. The emitter resistors $R_{46}$ and $R_{47}$ are introduced to give local degeneration. They provide a high impedence input to the compound transistor, which permits a voltage drive source, and also linearize the output current. Further, they improve the frequency response of the output stage. The load consists of the 30$\mu$H inductance in series with $R_{48}$, a $\frac{1}{2}\Omega$ resistor which monitors the current. The voltage across this resistor is attenuated (approximately by two) and is fed back to the voltage amplifier. The inductance is damped by $R_{59}$ to reduce ringing. To ensure the absence of spurious responses, it is necessary that the power resistors $R_{46}$, $R_{47}$, and $R_{48}$ are non-inductive. The phase splitter consists of two unity gain stages. The standing power dissipation in the transistors is reduced by the inclusion of resistors $R_{37}$ and $R_{40}$, which are by-passed by capacitors $C_6$ and $C_7$. The diodes $MR_3$, $MR_4$, and $MR_5$ are inserted for thermal compensation. When the appropriate phase splitter transistor switches off, the diode will become high impedance and the base of the 'power transistor' will be 'floating'. To overcome this, resistor $R_{43}$ maintains the diodes forward biased under all conditions. As a precaution against the diodes going open-circuit, they are by-passed by resistors $R_{42}$ and $R_{44}$. Additionally, the network $MRZ_1$, $MRZ_2$, $R_{45}$, $MR_6$ and $MR_7$ ensures that, if one of the diodes should become open-circuit, or if the input signal to the amplifier is too large (i.e. in excess of $\pm\frac{1}{2}$V) then the voltage signal developed at the base of the 'power transistor' will be limited to the Zener voltage. Without the diodes $MR_6$ and $MR_7$, the Zener capacitance, which is high when the Zener is non-conducting, would be in parallel with the voltage source to the output stage. This would seriously reduce the h.f. performance, and cannot be tolerated. Thus, it can be seen that the apparent complexity of the phase splitter stage, is all concerned with protection of the output stage against overload. This is necessary, since the power stage by itself, cannot be so protected, and

TABLE 1
Abridged Specification

| | | |
|---|---|---|
| Input | = | $\pm$0·5V |
| Output | = | $\pm$2A |
| Input impedance | > | 250$\Omega$ |
| Load | = | 30$\mu$H coil |
| Frequency response | = | 0 to 100kc/s |
| Temperature stability | = | Better than 1% of f.s.d. |

* *Formerly Kelvin Hughes Division, S. Smith & Sons (England) Ltd.*

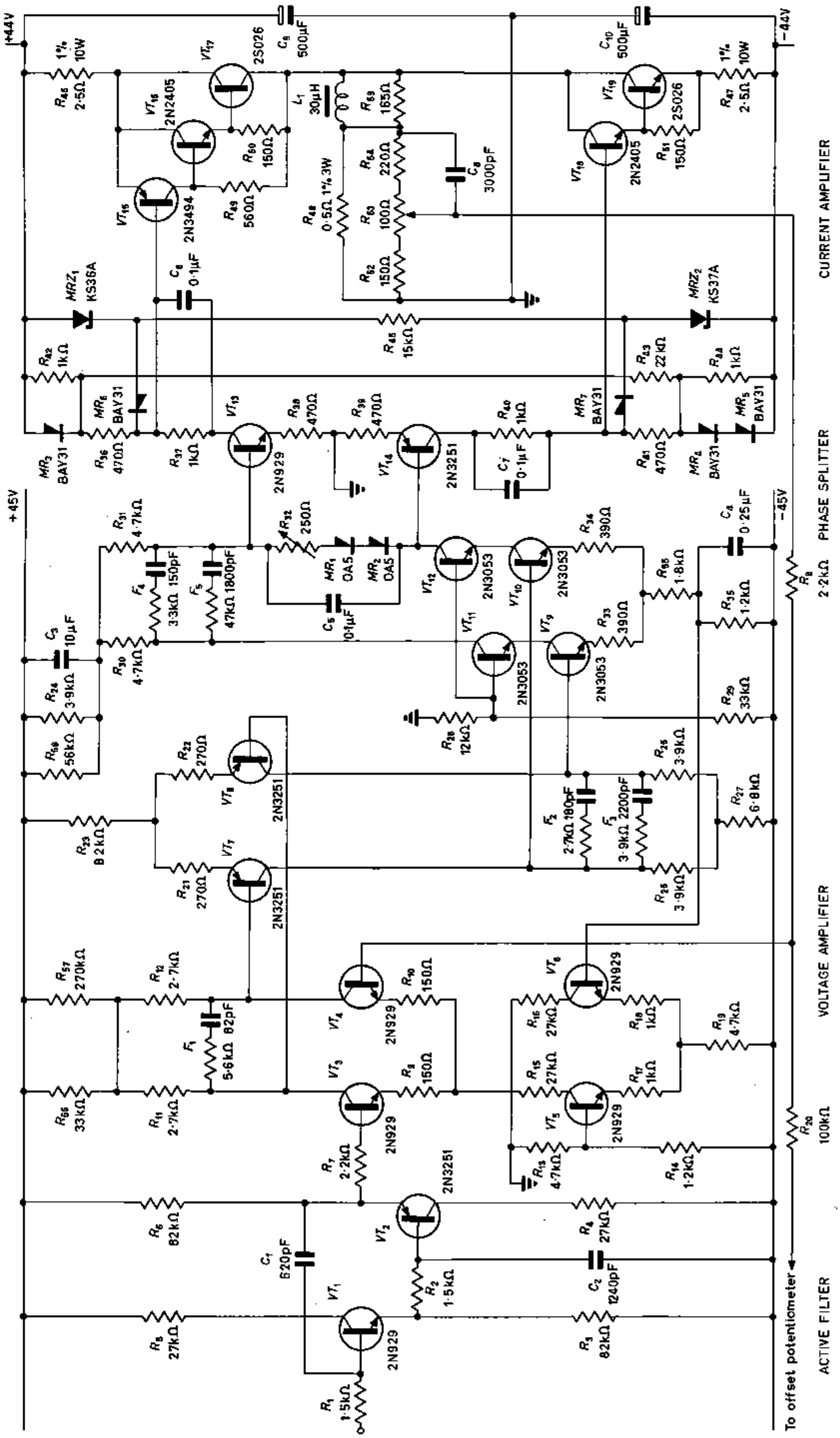



Fig. 2. The complete amplifier circuit

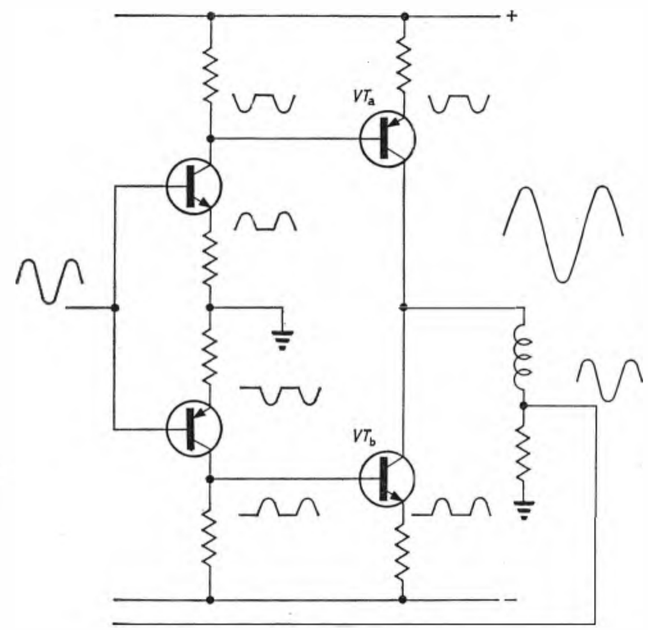



Fig. 3. Structure of output stage

the fail-safe mechanisms must be inserted elsewhere.

Some care was taken with the design of the voltage amplifier to ensure a high frequency response coupled with good stability. The output signal was designed to be $\pm5$V maximum. This was fixed by consideration of the maximum reverse emitter base voltage of the transistors in the phase splitter stage, since these are back biased by part of the input signal (see Fig. 3). Thus with an input of $\pm0{\cdot}5$V the closed loop gain is fixed at 20dB. The open loop gain consisted of a compromise between conflicting requirements; the amount of feedback should be high to ensure good linearity, and it should be low to maintain high frequency stability. A reasonable compromise was to select an open-loop gain of 60dB at d.c. Since any high frequency roll-off would be based on this figure of 60dB, it became necessary to ensure that the open-loop gain, itself, was sensibly stable and reproducible. That is to say, the gain should be independent of a change of active element. Further, it is important that the response be maintained well beyond the operating frequency range in order that the roll-off networks may operate without additional phase shift from the amplifier. For these reasons a three-stage amplifier was used with low output loads, to reduce the Miller effect, and a large amount of local feedback to give reproducible stage gains.

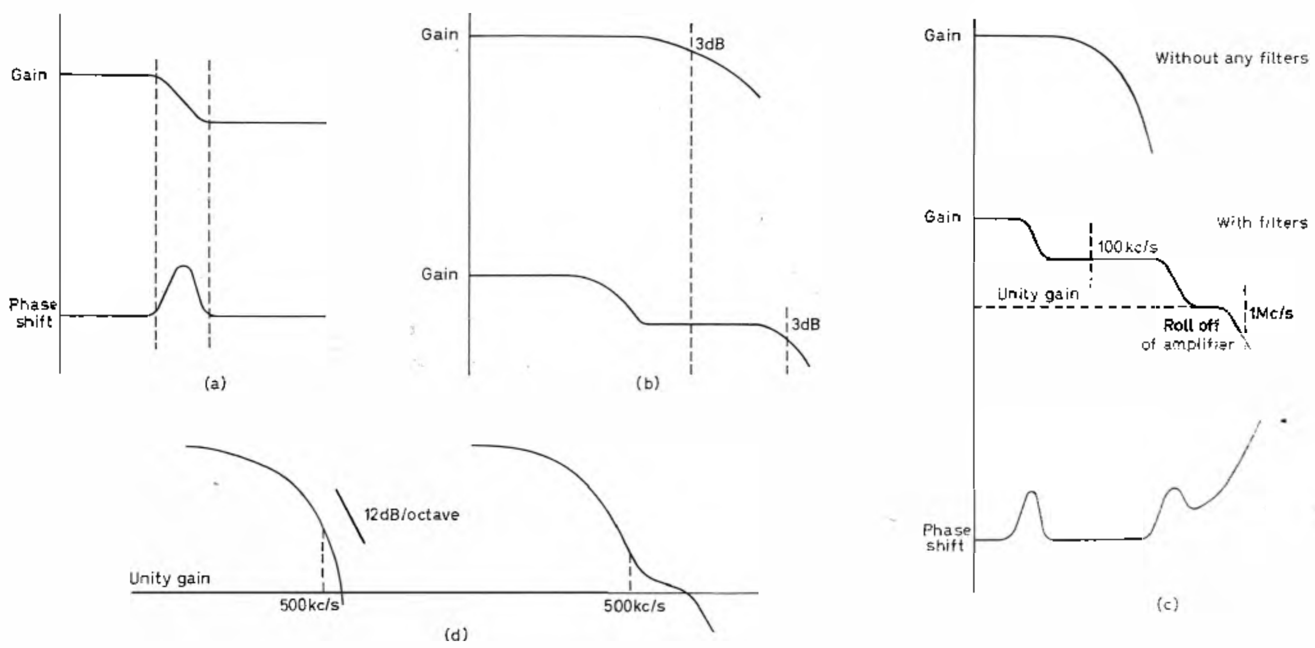
There are two feedback loops in the voltage amplifier. The first controls the overall gain and is of the Slaughter type, i.e. the feedback is in phase with the input and is returned to the base of the 'free' transistor in the first long tailed pair. The second loop reduces the common mode gain. Their modes of operation have been discussed elsewhere[1]. In the absence of input signal, the output may be set to zero (or varied over the whole range of operation) by means of the offset control. This is part of the power supply and is connected to the 'free' transistor of the first long tailed pair by $R_{20}$. This method of zero-adjustment was adopted since it does not interfere with the gain of the first stage, unlike the more common methods in use.

The first stage, consisting of transistors $VT_3$ and $VT_4$ has a stage gain of about 16dB. It is fed from a constant current source, transistor $VT_5$, which is itself part of a long tailed pair. The other transistor of this pair controls the current output from transistor $VT_5$ by means of common mode feedback from the base of resistor $R_{35}$. This feedback is for d.c. purposes only and is a.c. decoupled by $C_4$. Resistors $R_{17}$ and $R_{18}$ stabilize the feedback while resistors $R_{15}$ and $R_{16}$ are inserted merely to reduce the voltage across transistors $VT_5$ and $VT_6$. The output of the first stage is fed to the second stage ($VT_7$ and $VT_8$), which in turn feeds the third stage ($VT_9$ and $VT_{10}$), both having stage gains of about 22dB. The last stage has a cascode configuration, incorporating transistors $VT_{11}$ and $VT_{12}$, to improve the frequency performance; and diodes $MR_1$, $MR_2$ and resistor $R_{22}$ are inserted at the output to control and stabilize the degree of forward bias in the power stages.

### Filter

A simple negative feedback amplifier may be defined as being non-oscillatory if the phase shift at the output

Fig. 4. Gain and phase shift characteristics



changes by less than 180° when the gain is in excess of unity. Now, a phase shift of 180° is equivalent to a 12dB/octave slope of the open-loop response curve, and this curve may be used as a measure of stability. Since the amplifier (in this case consisting of six stages) may have a frequency response rolling off at something far greater than 12dB/octave, it is necessary to artificially reduce the gain to less than unity before the phase shift of the amplifier (plus any filter) exceeds 180°. This is accomplished by means of filter units ($F_1$ to $F_5$ in the circuit diagram) which have the property of attenuating the open-loop gain without additional phase shift at high frequencies. This is shown in Fig. 4(a). They have an additional advantage. Since they reduce the gain of a stage, they extend the frequency response, albeit at an attenuated level, which is shown in Fig. 4(b). This permits still further filtering to be used over the linear part of the response curve. It is for this latter reason that the filtering has been shared over each of the three voltage gain stages.

The amplifier in question has an open loop gain of 60dB at d.c., and a closed loop gain of 20dB. Thus, it is desirable that the filter provide an attenuation of 40dB before the phase shift of the amplifier itself becomes appreciable. It was decided than an attenuation of 20dB can be tolerated within the operating frequency range (i.e. there is only 20dB feedback at 100kc/s) and the remaining gain attenuated at as low a frequency as possible outside this range. The approach is shown in Fig. 4(c). The first attenuation is provided by two 10dB filters $F_3$ and $F_5$ which operate approximately over the band 7kc/s to 20kc/s. The rest is provided by three 6dB filters, $F_1$, $F_2$ and $F_4$, which each have a centre frequency of 250kc/s. The evaluation of these filters is given in the Appendix.

As an additional safeguard, some phase advance is introduced near the unity gain frequency, by increasing the amount of feedback at this point. The capacitor $C_8$ provides a 6dB lift with the 3dB point at about 500kc/s. This has the effect of reducing the slope (as shown in Fig. 4(d) which enhances the stability at this point.

The active filter which consists of transistors $VT_1$ and $VT_2$ plus their associated circuits, introduces a Butterworth type low-pass filter with a 12dB/octave roll off and a 3dB down point[2] at 100kc/s. Since the output load (the inductance) will have a voltage swing rising at only 6dB/octave with increase of frequency, there will be a net drop in output voltage above 100kc/s. The filter is necessary to prevent saturation of the output stage, with high frequency inputs, since saturation under these conditions can induce high frequency instability. The timing networks are $R_1C_1$ and $R_2C_2$, and the capacitors are evaluated by the following formulae: $C_1 = 1/(\sqrt{2}\,.\,2\pi f_o\,.\,R_1)$; and $C_2 = 2C_1$, where $f_o$ is the 3dB down frequency.

### Thermal Compensation

In order to avoid crossover distortion, the output stage is forward biased with 50 to 100mA flowing through both output transistors in the absence of any input signal. In this condition it is possible for thermal runaway to occur, and the problem is overcome by means of a thermal feedback loop. This loop generally consists of a forward biased diode in parallel with the transistor emitter base diode connected thermally with the collector. Thus, without the diode, if the transistor heats up, the emitter base voltage will drop and, with a constant voltage source, the output current will increase giving more heat to the transistor. With the diode in circuit, heating the transistor will equally heat the diode, causing an equivalent drop in the forward diode voltage, which will compensate for the emitter base diode change, and the collector current remains constant. This system has been used where diodes $MR_1$ and $MR_2$ compensate for transistors $VT_{13}$ and $VT_{14}$. Since control of the forward bias is desired, germanium diodes are used, and the voltage drop across $R_{22}$ gives the required regulation of the output current. Examination of the pnp compound output transistor shows that the controlling base emitter voltage is that of transistor $VT_{15}$. Since this is running at very low power levels, the compensating diode ($MR_3$) need not be attached to any heat sink, but must merely respond to ambient changes. On the other hand, $MR_4$ and $MR_5$ (compensating $VT_{18}$ and $VT_{19}$) should be attached to the output heat sink. However, this has not been done, since the pnp transistor being fully compensated, any change of current through transistor $VT_{19}$ will constitute an offset signal and will be reduced by the feedback of the whole amplifier. Thus, the inconvenience of diode connecting wires trailing from the circuit board to the heat sinks has been avoided, with no loss of thermal stability. The voltage amplifier has the first stage enclosed in a heat sink, has common mode as well as overall feedback, and, as such, should have low thermal drift. The active filter, being outside the feedback loop becomes a controlling factor in overall drift. For this reason, $VT_1$ and $VT_2$ are enclosed in a heat sink, and run at fairly low and equal collector currents. In this way, it is hoped that the base emitter variations of both transistors will cancel to give adequate thermal drift characteristics.

### D.C. Supplies

The maximum peak voltage developed across the load is given by $V_{max} = I_{max}\,.\,2\pi f_{max}\,.\,L$, where $I$ is the current passing through the load, and $L$ the load inductance. Substituting values from Table 1 gives $V_{max} = 36$V. From this figure the voltage of the output stage power supply is fixed at 44V. Examination of phase splitter plus output stage shows that variation of the power supply voltage will have no effect on the output signal. Thus, the high current power supply need not be well stabilized or regulated. On the other hand, the low current supply to the voltage amplifier requires high stibilization and regulation. Further, it is convenient to use it to provide a measure of control to the high power supply. For this latter reason, the system was made short-circuit safe using a high impedance source driving a shunt regulator. While this becomes inefficient with high currents, it is most convenient at low levels[3,4]. The basis of the circuit is shown in Fig. 5, and the full circuit diagram in Fig. 6. While designed on conventional lines, it has the advantage that the 'overvoltage' rail for each half (point $B$ in Fig. 5) is returned to the other half of the power supply providing a well stabilized source for the comparator rail.

Fig. 5. Basis of power supply

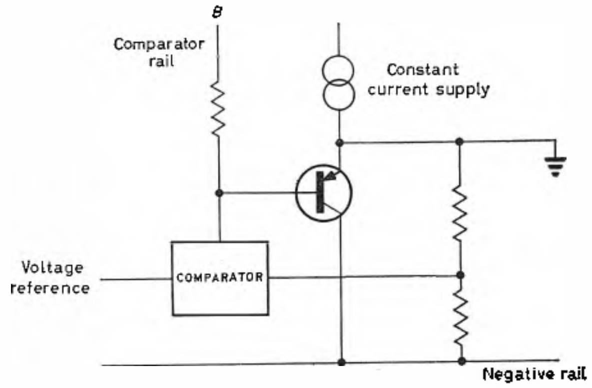


**TABLE 2**
**Performance of Amplifier**

| | | |
|---|---|---|
| Input | = | ±0·5V |
| Output | = | ±2A inductive load |
| Input impedance | > | 1·5kΩ |
| Frequency response | | |
| D.C. to 70kc/s | = | 1dB down |
| D.C. to 100kc/s | = | 3dB down |
| Linearity of Gain | = | Better than 0·3% of f.s.d. |
| Temperature stability (25 to 55°C) | = | Better than 0·5% of f.s.d. |
| Bias control | — | Can operate over full range of ±0·5V |
| Overload | — | 300 per cent will cause no permanent distortion |

Diodes $MR_1$ to $MR_4$ are inserted to ensure that, under all conditions, the base emitter diodes of transistors $VT_9$ and $VT_4$ will not have reverse voltage breakdown, while diodes $MR_5$ and $MR_6$ ensure that the collector base diodes of the power transistors will never conduct in the forward direction. The potentiometer $R_4$ provides the offset voltage for the amplifier.

**Results**

The performance of the amplifier is shown in Figs. 7, 8, and 9, and the results tabulated in Table. 2. The shape of the closed-loop response exactly follows that of the active filter, and the effect of the roll off filters on the open-loop response can clearly be seen from Fig. 8. The amplifier was tested over a 30°C temperature range, with the input short-circuited. The active filter alone has a drift of 2·4mV and, measuring across the monitor resistor

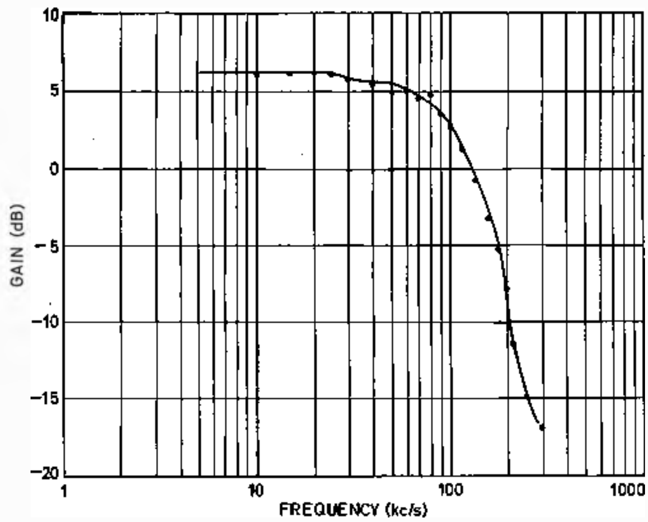
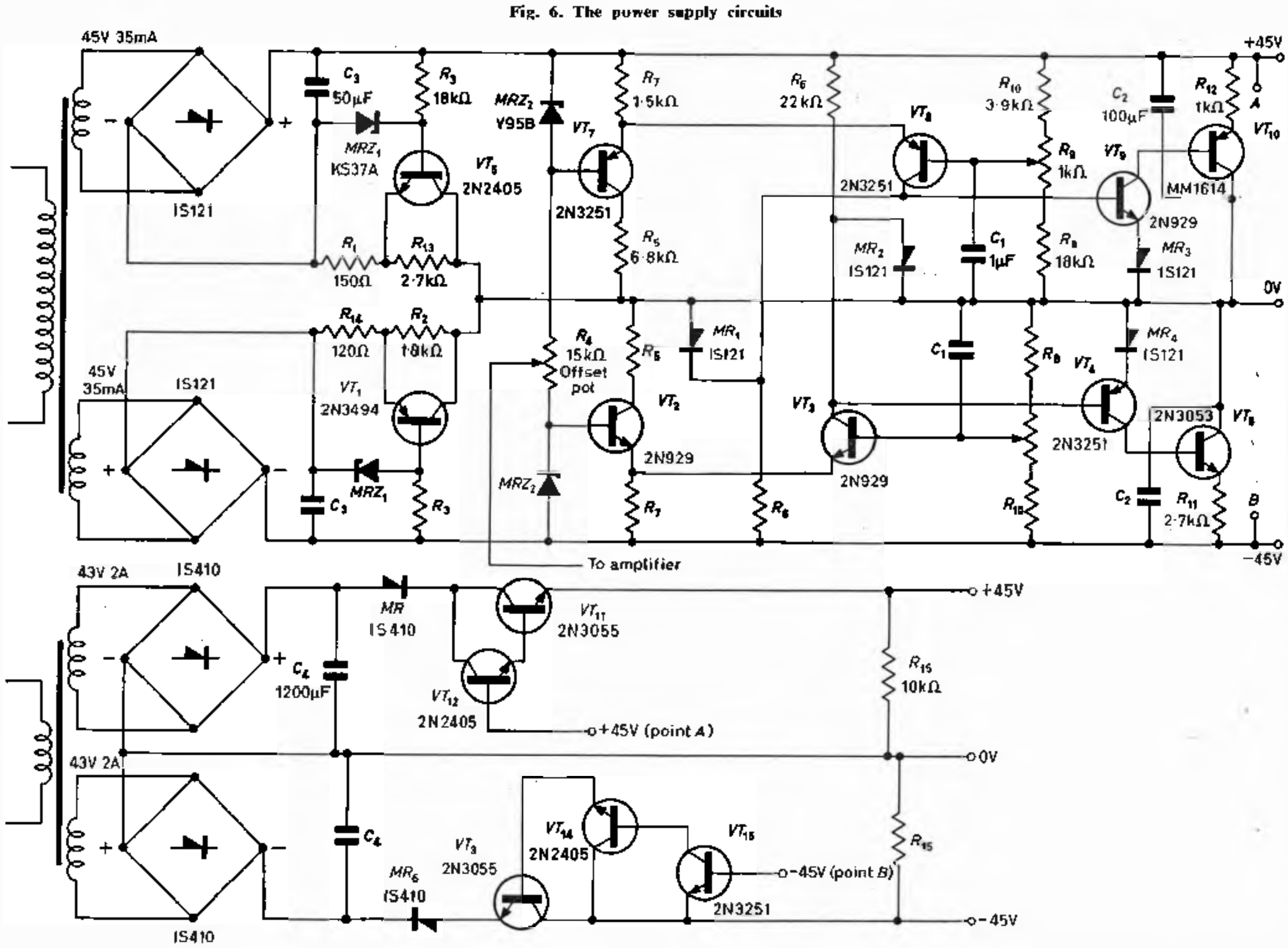



**Fig. 7. Closed loop response**

at the output, the drift was measured as 3·0mV. Since the voltage gain of the amplifier is approximately 2, the drift at the output due to the amplifier alone is $2 \times 2{\cdot}4 - 3{\cdot}0 = 1{\cdot}2$mV, i.e., the drift referred to the input = 30$\mu$V/°C which is more than adequate. Better results could probably be obtained with more common mode feedback but, with this amplifier, if lower drift is required, then more attention should be paid to the active filter components. Again, with the input short-circuited, the change in current in the output transistors (set to

**Fig. 6. The power supply circuits**



100mA at 25°C) was less than 1/2 per cent over the 30°C temperature range, which shows the efficacy of the thermal compensation. While the offset control would permit the output to be set to any level, the extreme positions (i.e., at $\pm 2A$) can only be maintained for a few minutes, since the heat sinks of the output transistors are only cooled by natural convection. With forced air cooling, the period could be extended. No difficulty was experienced with the power supplies. The stabilized units have a ripple content of 1mV r.m.s. and change less than 10mV d.c. for all operating conditions. The low power supplies can be short-circuited with no permanent damage, although this cannot be done to the high current outputs.

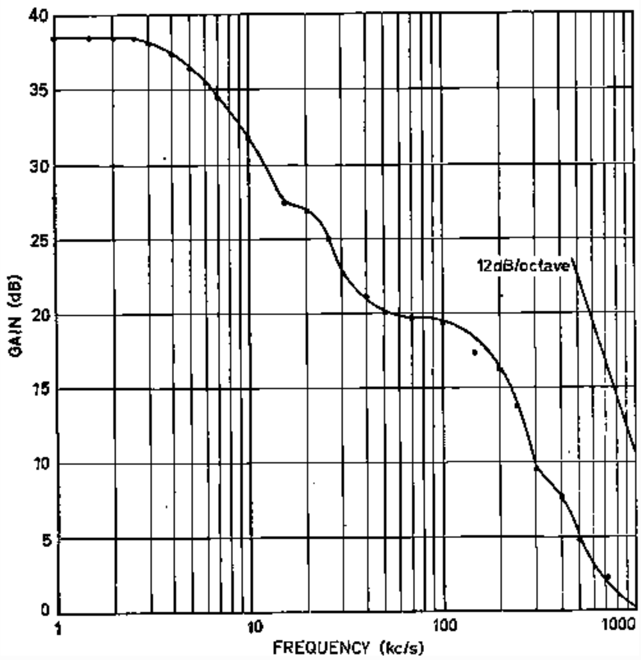



Fig. 8. Open loop response

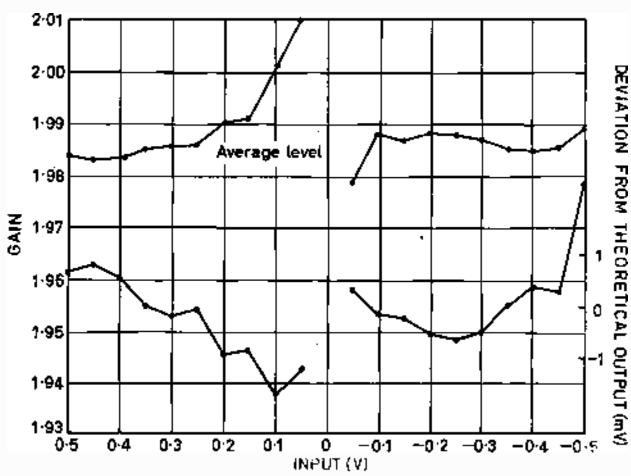



Fig. 9. Input voltage-gain characteristics

## Conclusion

It has been shown that a wideband, high power amplifier can be used with an inductive load, without resort to exotic active elements. Although care was taken with the high frequency stability, little effort was taken with the d.c. performance, which would suggest that improvements can readily be obtained in the thermal drift characteristics, should this prove necessary.


**Acknowledgments**

The above work has been carried out in the Kelvin Hughes Division of S. Smith and Sons, for a piece of instrumentation of another Division, and the author would like to thank those responsible, for permission to publish this work.


## Appendix

The analysis of the filter is made using a star-delta transformation (see Fig. 10), where the current generators provide out of phase signals, representing the transistors of a long tailed pair.

Let the d.c. load $= R$
the components of the filter $= r$, and $C$
the required attenuation $= 1/n$
and the phase shift at any frequency $= \beta$

then:

$$Z = \frac{R(1 + j\omega Cr)/j\omega C}{2R + \dfrac{(1 + j\omega Cr)}{j\omega C}}$$

$$= \frac{R(1 + j\omega Cr)}{1 + j\omega C(r + 2R)} \quad \ldots\ldots\ldots\ldots\ (1)$$

This represents the effective output load since there will be no signal across the impedance $z$.

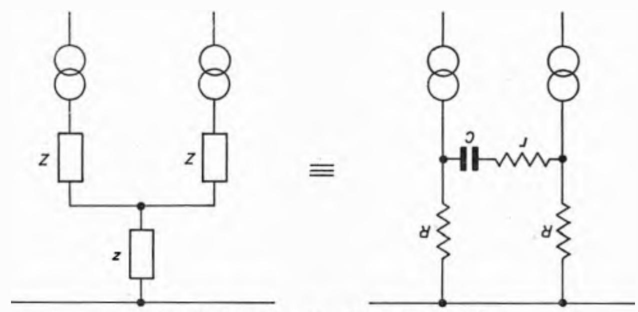



Fig. 10. Star-delta transformation of filter

At infinite frequency

$$Z = R \cdot \frac{r}{r + 2R} = R/n$$

$$\therefore nr = r + 2R$$

$$r = \frac{2}{n - 1} \cdot R \quad \ldots\ldots\ldots\ldots\ (2)$$

Putting $\omega CR = p$

and $\dfrac{2}{n-1} = m$

$$\text{then } Z = \frac{R(1 + jmp)}{1 + jmnp}$$

rationalizing and simplifying

$$= \frac{R(1 + m^2p^2n) + R \cdot jmp(1 - n)}{1 + (mnp)^2} \quad \ldots\ldots\ (3)$$

Now, at the 3dB point $|Z|^2 = R^2/_2$

But

$$|Z|^2 = \frac{R^2(1 + m^2p^2n)^2 + R^2m^2p^2(1 - n)^2}{(1 + m^2n^2p^2)^2}$$

i.e. $(1 + m^2n^2p^2)^2 = 2(1 + m^2p^2n)^2 + 2m^2p^2(1 - n)^2$

which gives $m^4n^2(n^2 - 2)p^4 - 2m^2p^2 - 1 = 0$

solving the quadratic, and simplifying:

$$p^2 = \frac{1}{m^2(n^2 - 2)}$$

or:

$$p = \frac{1}{m\sqrt{(n^2 - 2)}} \quad \ldots\ldots\ldots\ldots\ldots\ (4)$$

$$= \frac{(n-1)}{2\sqrt{(n^2-2)}}$$

thus:

$$C = \frac{(n-1)}{2R \, . \, \omega_{3dB} \, . \, \sqrt{(n^2-2)}} \quad \ldots\ldots\ldots\ldots (5)$$

Equations (2) and (5) give the required component values for a 3dB down point at angular frequency $\omega_{3dB}$.

The maximum phase shift of the network is derived from equation (3).

$$\tan\beta = \frac{mp(1-n)}{1+m^2p^2n}$$

But $m = 2/(n-1)$

$$\tan\beta = -\frac{2p}{1+m^2p^2n} \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

The negative sign merely denotes a phase lag and will now be ignored!

$$\frac{d(\tan\beta)}{dp} = \frac{2(1+m^2p^2n) - 4m^2p^2n}{(1+m^2p^2n)^2}$$

for maximum value of $\beta$ (which is equivalent to maximum value of $\tan\beta$)

$$2(1+m^2p^2n)^2 = 4m^2p^2n$$

$$m^2p^2n = 1$$

$$p = 1/m\sqrt{n}$$

substituting in equation (6):

$$\tan\beta_{max} = \frac{2 \, . \, 1/m\sqrt{n}}{1+m^2n \, . \, 1/m^2n} = 1/m\sqrt{n}$$

i.e.

$$\beta_{max} = \tan^{-1}\frac{(n-1)}{2\sqrt{n}} \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

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## A British Satellite Communication Station

A British satellite communications earth terminal station for the American Apollo 'man-on-the-moon' programme is to be designed, built and installed on Ascension Island by The Marconi Co. Ltd for Cable and Wireless Ltd.

Two contracts have been awarded to Cable and Wireless by the National Aeronautics and Space Administration in Washington, for the provision of key communications systems, on Ascension Island and in Bermuda, for the Apollo project, designed to put an American on the moon before 1970. The Ascension Island satellite station, will communicate through a synchronous satellite similar to Early Bird, with the Goddard Space Flight Centre in Maryland, via the Andover, Maine, Earth Station. Marconi's are also to equip a high power, high frequency communications system in Bermuda, linking the Apollo ship in mid-Atlantic with Bermuda and Goddard. Both stations have to be operated in less than a year, to meet the very tight Apollo schedule.

The earth terminal station on Ascension Island will be linked to the Andover Earth Station via a synchronous satellite in orbit over West Africa. Stations in the Canary Islands, and on an Apollo ship in mid-Atlantic will also use this satellite. The equipment will be fully air transportable, as a number of major sub-units, to facilitate rapid installation.

The system uses a 42ft diameter, steerable, parabolic dish aerial with a focal length of 12ft, and with a centre area accuracy of 0·03in. Even at the edge, the aerial will be within 0·075in of the ideal paraboloid.

The dish surface will be made up of 2in thick aluminium honeycomb sandwich panels, mounted on an aluminium backing framework. This very rigid structure is attached to an elevation mounting, which is in turn mounted on a platform which can turn through 370° in azimuth. The majority of the transmitter and receiver equipment will be mounted on this platform, and both the received and the transmitted signals will pass to the ground installation through coaxial cables at the intermediate frequency of 70Mc/s.

The transmitter will operate in the 6 000Mc/s frequency band, with a final output power in the region of 15kW. The output klystron, mounted on the back of the dish structure, will feed power through a waveguide system to a horn mounted in the surface of the dish. A Cassegrain system will be used, with a small hyperbolic sub-reflector at the focus of the main dish.

The receiver, operating in the 4 000Mc/s frequency band, will use the same Cassegrain feed system and circular waveguide horn. It will also be mounted directly on the back of the dish aerial. A cryogenic parametric amplifier, operating at −253°C in a helium system will be used, to reduce the noise temperature of the receiving system. Extremely high amplification is necessary and the small amount of noise produced by receivers at normal temperatures would swamp the tiny signals received from the satellite.

The accuracy of the aerial system, with its very low sidelobe response and low temperature receiver, ensure a low ratio of noise to the required signals, even at amplifications in the region of 200 000 times. This amplification is essential to ensure the reliability of communications under the worst conditions.

Power supplies and all major elements of both the transmitting and receiving system are fully duplicated, to ensure maximum reliability. The standby equipment is maintained in a fully prepared condition, and switching to standby for any unit can be effected within a fifth of a second.

The aerial attitude control system will be of the auto-follow type.

The high frequency communications station at Bermuda is designed to carry two high speed data transmission circuits, each with a maximum capacity of 2 400 bits, a speech channel and three teleprinter channels to the Goddard Space Flight Centre, from Bermuda and from the NASA Apollo ship, which is normally located some 1 000 miles south east of Bermuda.

Five of the latest Marconi self-tuning 30kW independent sideband transmitters type H1200, will be installed at the Cable and Wireless transmitting station at St. George's, to operate with five Collins rotatable log periodic, wideband aerials. Each of two of the transmitters will carry the two data channels, using separate frequencies, to ensure that the effects of noise and fading are minimized. A further two transmitters will each transmit the speech channel and the three frequency modulated, voice frequency teleprinter channels, again providing frequency diversity. The fifth transmitter will act as a standby to cover frequency changes and possible faults.

At the Devonshire receiving station, three dual path independent sideband, solid state, electronically tuned receivers, type H2102 will be provided, with two more rotatable log periodic aerials. The third receiver will again act as an emergency standby and an additional aerial will cover emergencies at either the transmitting or receiving stations. Telephone terminal units, type HW23 will be provided together with twelve channel, frequency modulated, voice frequency equipments, type H5000/5001, which will provide three teleprinter channels and an order wire facility over the telephone cable from Bermuda to New York.

A comprehensive, high stability drive and frequency control system will be provided for both transmitting and receiving stations, based on a Marconi frequency standard 1Mc/s oscillator, with a short term stability better than ±3 parts in $10^{10}$.

# A Slope-Quantized Binary Pulse Code Modulation

By M. N. Faruqui*

*A new technique of coding signal slopes with uni-digit 1/0 pulses, using a special feedback circuit and a threshold in the coder-comparator, has been developed. The overall performance of the system has been found to be better than those of delta and delta-sigma modulation systems in many respects. The dynamic range of the system has been extended through the use of an instantaneous compandor but the system uses only simple circuits. The characteristics of the system are discussed and compared, and it is shown that this system at 40kc/s p.r.f. is equivalent to the familiar six-digit p.c.m. in all respects.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

DIGITAL transmission of message signals has many advantages over analogue transmission, and for this purpose, various analogue-to-digital conversion techniques have been developed. The familiar amplitude-quantized multi-digit pulse-code modulation[1] (AQ-PCM) systems use complex circuits for coding, and their coding efficiencies for smaller numbers of digits are poor. The simpler coding techniques, e.g., the delta-modulation[2] ($\Delta$-M) and $\Delta$-$\Sigma$M systems[3], where binary positive and negative pulses approximate the analogue voltage using a single digit code, require bandwidths larger than those of amplitude quantized p.c.m. for an equivalent performance. Moreover, the $\Delta$-M system has poorer signal-to-noise ratios and smaller outputs at higher signal frequencies, and the $\Delta$-$\Sigma$M equalizes these characteristics at the cost of lower overall signal-to-noise ratio.

In an attempt to improve the coding efficiency of the simpler coded systems, a uni-digit binary slope-quantized system has been developed where the frequency response characteristics are improved by using a somewhat modified technique of quantizing the signal slopes, has been developed. In a similar manner as used in amplitude quantized p.c.m., a compandored slope quantized p.c.m. system has been found to have an almost uniform signal-to-noise ratio for wide dynamic ranges of the input, although some earlier authors[4] have said to the contrary. The results obtained here show that uni-digit slope quantized p.c.m., using 1/0 pulses only, has an overall performance equivalent to that of multi-digit amplitude quantized p.c.m. using approximately the same video bandwidth.

### Theory of Binary Slope Quantized P.C.M.

The approximation of waveforms is usually done by using $\sin x/x$ type of interpolating functions, but Gabor[5] has shown in his generalized sampling theorem that other interpolating functions, like the exponentially decaying ones, are equally satisfactory for representing a signal. A band-limited signal $f(t)$ may, then, be represented as:

$$f(t) \simeq \sum_{n} C_n r(t - t_n) = f'(t) \quad \ldots\ldots\ldots\ldots (1)$$

where $f'(t)$ represents the approximated value of $f(t)$, $r(t - t_n)$ is the response of the interpolating filter to an impulse occurring at $t = t_n$, and $C_n$'s are the sample values obtained by passing $f(t)$ through a network whose frequency domain characteristics are complementary to those of the interpolating filter. If the interpolating function is the type of response obtained from a partial integrator, shown in Fig. 1(a), then $f(t)$ has to be passed through a complementary differentiator before sampling and quantization are done. By making the sampling intervals much smaller than the Nyquist interval of $1/2\omega_0$ and by using a matching $r(t)$, a further approximation for such signals, as are made unidirectional in the process of

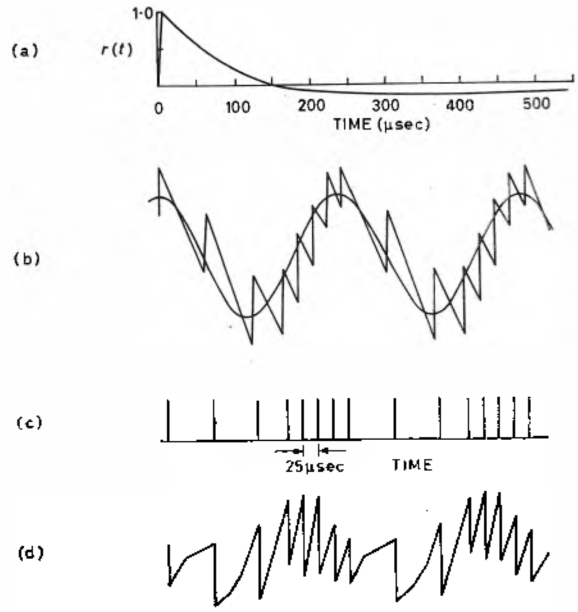


**Fig. 1. Approximations of sinusoidal signal**
*(a) Response of the feedback network to a pulse of 5μsec duration*
*(b) Oscillogram of a reconstructed 3·33kc/s signal (with p.r.f. of 40kc/s) along with the sinusoidal signal of the same frequency superimposed*
*(c) Transmitter output*
*(d) The error signal*

sampling, may be obtained by using quantized values of $C_n$ only, thus:

$$f'(t) \simeq \sum_{n} C_o \,.\, r(t - t_n) \quad \ldots\ldots\ldots\ldots (2)$$

where:

$$C_o = \begin{cases} 1 \ , \text{ if } |C_n| \geqslant |\text{r.m.s. } C_n| \\ 0 \ , \text{ if } |C_n| < |\text{r.m.s. } C_n| \end{cases}$$

The signal coding into a sequence of 1/0 pulses, using the above principles, may be done in a simple circuit shown in Fig. 2, where the error $E(t)$ is sampled and quantized in the comparator with an optimum reference. The feedback network builds up $B(t)$ from 1/0 output pulses and again compares it with $f(t)$, thereby producing $E(t)$. The reference of the quantizer and the interpolating response $r(t)$ (of the feedback network) are so adjusted as to give the minimum error in the reconstructed signal $f'(t)$ at the receiver output. The use of feedback in the

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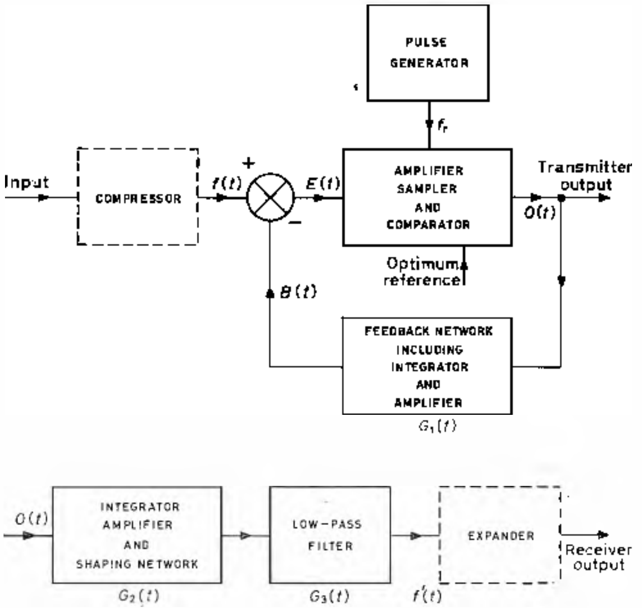



**Fig. 2. Arrangement of transmitter and receiver of binary SQ-PCM system**

coding process provides the partial differentiation of the input as required by equation (1), and improves the approximation of $f(t)$ by a continuous comparison between $f(t)$ and $B(t)$. Oscillograms of an approximated signal $f'(t)$, the coder output $O(t)$, the error $E(t)$ and the interpolating function $r(t)$ are shown in Fig. 1, where the correlation between the occurrence of 1/0 pulses and slopes of the smoothed curve is evident. In practice the feedback network response is so adjusted that the negative slopes of $f(t)$ are matched by the cumulative negative slopes of $B(t)$ when no pulses occur at the output. The composite modulator, then, operates on the successive differences between $f(t)$ and $B(t)$, and the successive approximations match the slope of $f(t)$, thus, giving rise to slope quantized p.c.m.

The coding process discussed above is somewhat similar to that used in $\Delta$-M and $\Delta$-$\Sigma$M, but the use of a definite threshold in the quantizer and a matching response $r(t)$ have been found to improve the overall characteristics of slope quantized p.c.m., even with simplified circuits giving only 1/0 pulses at the coder output. Moreover, the coding with $+1/-1$ pulses without any threshold in the comparator, as is done in $\Delta$-M and $\Delta$-$\Sigma$M, produces some extraneous pulses, which do not contribute to the minimization of the error waveform, but tend to produce some extra error in $f(t)$.

### Quantizing Noise

As in all approximations, the residual error in $f'(t)$ will be somewhat random[6] when the input signal is semi-random, e.g., speech. For sinusoidal input signals, the error waveform consists of approximate triangular pulses, shown in Fig. 1. The error amplitudes will vary between $\pm V_e$, where $V_o$ is the comparator output $O(t)$ (the gain of $G_1(t)$ is normalized to give the same peak value of $B(t)$). If it is assumed that all error amplitudes are equally probable between $\pm V_e$, then the error entropy is maximum and the signal-to-noise ratio at the receiver output, given by equation (10) in the Appendix, is minimum. The use of an optimum threshold and a matching response of the feedback network, however, reduce the error entropy by decreasing the number of large amplitude pulses. This gives rise to a peaky probability distribution for the error voltages, and the optimized distribution would approach a truncated normal distribution[7] for complex $f(t)$. The minimum error power is then calculated by assuming the average crest factor of four (as used for random noise and speech signals), and the signal-to-noise ratio obtainable for $f(t)$ is now given by (referring to the discussion in the Appendix):

$$[S/N] = 0.5\,\frac{k\,f_r^{3/2}}{f_c^{1/2}\,.\,f_m}\ ;\quad k \leqslant 1 \quad \ldots\ldots\ldots (3)$$

It is now seen that the system signal-to-noise ratio decreases directly with the amplitude ($\equiv K$) of $f(t)$, as shown in Fig. 3, and inversely with signal frequency $f_m$. The improvement with p.r.f. would be 9dB/octave. For semi-random signals like speech, the signal-to-noise ratio would be less by about 9dB from the above optimum values[6].

As the signal-to-noise ratio of $f(t)$ decreases for lesser amplitudes of the signal, an instantaneous compandor may be used to equalize the ratio for the useful dynamic range, as has been used for amplitude quantized systems. From Mallinckrodts[8] discussion of the noise advantage obtained in an instantaneous logarithmic compandor, it may be shown that:

$$[S/N] = V_{it}/V_n \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

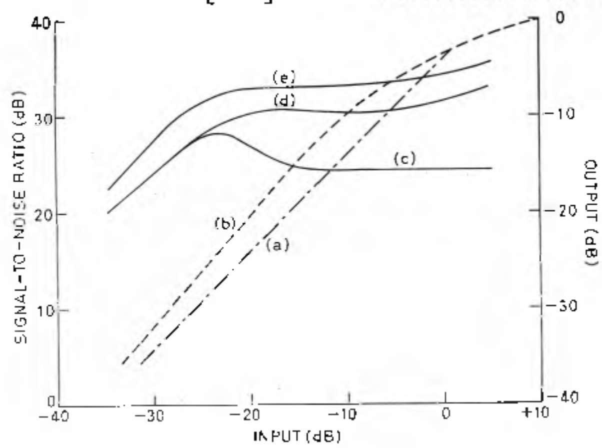
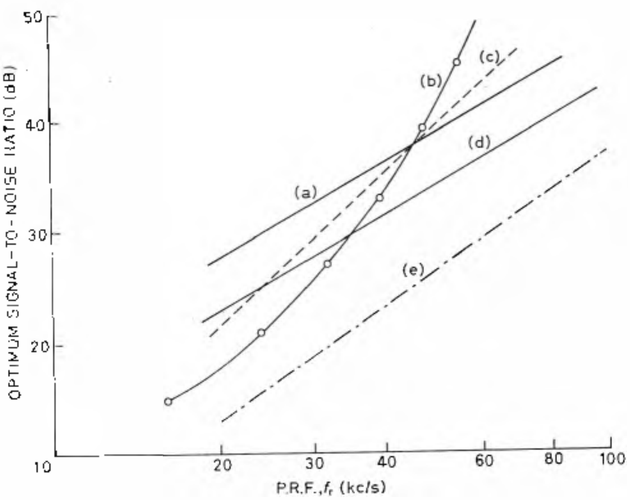



**Fig. 3. Theoretical signal-to-noise ratio of the SQ-PCM system**

(*a*) *Variation of signal-to-noise ratio with input for non-compandored system at 60kc/s p.r.f.*
(*b*) *Variation of output with input for the partial compandor*
(*c*) *Variation of signal-to-noise ratio with input for the linear compandor at 40kc/s p.r.f.*
(*d*) *Variation of signal-to-noise ratio with input for the partial compandor at 40kc/s*
(*e*) *Variation of signal-to-noise ratio with input for the partial compandor at 60kc/s*

**Fig. 4. Variation of optimum signal-to-noise ratio of different systems with p.r.f.**

(*a*) *Binary SQ-PCM*
(*b*) *AQ-PCM*
(*c*) *$\Delta$-M double integration*
(*d*) *$\Delta$-M single integration*
(*e*) *$\Delta$-$\Sigma$M*



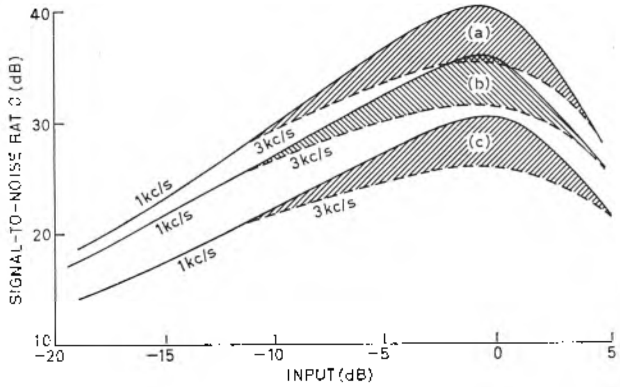



**Fig. 5. Variation of signal-to-noise ratio with input in the non-compandored SQ-PCM (the range of variation with $f_m$ is shown by the shaded area)**

*(a) 60kc/s p.r.f.*
*(b) 40kc/s p.r.f.*
*(c) 30kc/s p.r.f.*

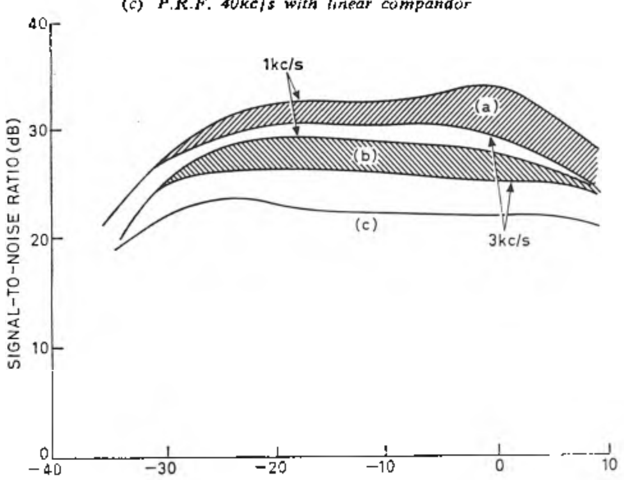
where $V_{it}$ is the input to the expander at the transition points of its characteristics, and $V_n$ is the total r.m.s. noise at the expander input due to the processing of signal and its harmonics through the coder. Using standard compressor-expander characteristics from Mallinckrodt's data and the signal-to-noise relation given in Fig. 5, the signal-to-noise ratio of the compandored system has been calculated and is shown in Fig. 3. For certain applications, e.g., in speech processing, it is not necessary to have the input-output characteristics of the compandor absolutely linear. Advantage may be taken of a partial compandor, whose characteristic is shown in curve (b) of Fig. 3, to obtain a better overall signal-to-noise ratio for the compandored system, also shown in Fig. 3.

**System Characteristics**

The experimental investigation of the proposed system has been done with the circuit shown in Fig. 2, where the feedback circuit in the quantizer includes an integrator with a break point of 800c/s and other equalizing networks. The response of the feedback network and the consequent signal built up $B(t)$, shown in Fig. 1, are such that the peak output flattens after 5 to 6 pulses and the peak-to-peak amplitude of $f'(t)$ is approximately $V_e \,.\, f_r/4f_m$ ($\equiv 3V_e$). To avoid overloading of the quantizer, then, the r.m.s. value of $f(t)$ has to be less than $0{\cdot}1\, V_e\, (f_r/f_m)$. The receiver with an audio bandwidth of 0·3 to

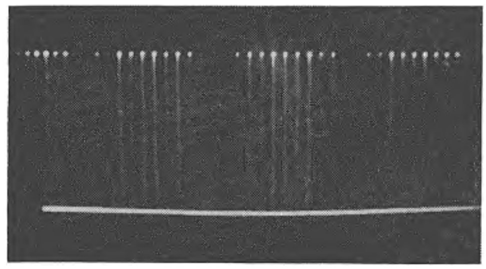
**Fig. 6. Variation of signal-to-noise ratio with input for the compandored SQ-PCM system (the range of variation with $f_m$ is shown by the shaded area)**

*(a) P.R.F. 60kc/s with partial compandor*
*(b) P.R.F. 40kc/s with partial compandor*
*(c) P.R.F. 40kc/s with linear compandor*

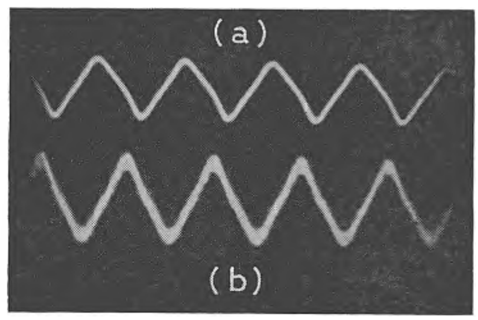



3·4kc/s is complementary to the transmitter, and uses local feedback to equalize the frequency response of $f'(t)$ and decrease the low frequency hum and noise. The overall system characteristics have been optimized for the useful p.r.f. range of 40 to 60kc/s.

The system characteristics, e.g., signal-to-noise ratio, frequency response, linearity etc., have been determined for the experimental set-up with and without compandors (both linear and partial) by using such test signals as sinusoidal, wideband and narrowband f.m., and noise. For sinusoidal signals, the variation of optimum signal-to-noise ratio with p.r.f. (without compandor) is shown in Fig. 4, along with the characteristics of other systems for com-



*Transmitted binary pulses*




*Reproduction of triangular waveform*
*(a) original input signal*
*(b) reproduced output signal*

parison. The variation of signal-to-noise ratio with input level for the non-compandored system is shown in Fig. 5, and that for the compandored system in Fig. 6. The gain/frequency response characteristics with and without compandor are shown in Fig. 7. The lower signal-to-noise ratio and output at higher frequencies for larger input levels in the non-compandored system may be attributed to the overloading of the quantizer, and the further decrease in the output at higher frequencies for the compandored system is due to the limited overall bandwith of the system.

The results of the noise test, using a half-band noise signal (300 to 1 800c/s) show that the signal-to-noise ratios are 27dB and 29dB for the non-compandored and 25dB and 27dB for the compandored system using p.r.f.'s of 40 and 60kc/s respectively. An f.m. signal, having a bandwidth similar to that of the noise test signal, gives (without compandor) an average signal-to-noise ratio of 28dB and 31dB for p.r.f.'s of 40 and 60kc/s respectively. A narrow-band f.m. signal occupying a band of about 400c/s gives (without compandor) an average ratio of 33 and 37dB at 40 and 60kc/s p.r.f. respectively. Live speech tests have been performed at 40 and 60kc/s p.r.f., with and without compandor, and the reproduction has been found to be excellent at 60kc/s p.r.f. The quality at

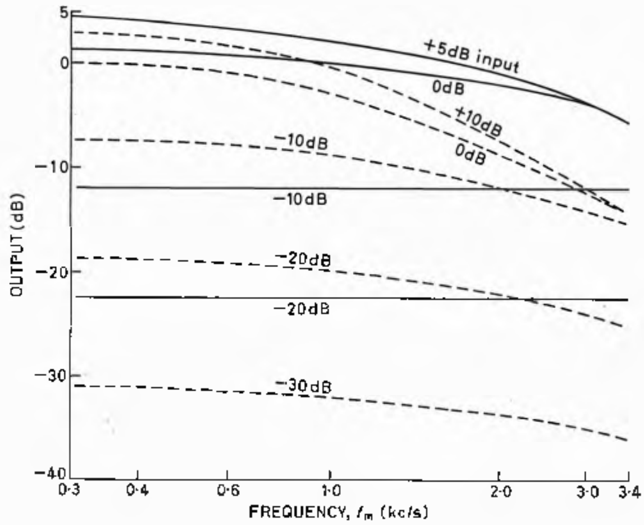



**Fig. 7. Variation of output with $f_m$ for different input levels in SQ-PCM at a 40kc/s p.r.f.**
(a) *Without compandor*
(b) *With partial compandor*

40kc/s p.r.f. with compandor is good but a little sandy in character; while without compandor, the speech is quite clear if it is allowed to overmodulate the system slightly.

An overall comparison of the different p.c.m. systems, with reference to the quantizing noise, dynamic range, and complex signal performance is shown in Table 1. For the transmission of slope quantized p.c.m. pulses, either as a video or an r.f. modulated signal, the power efficiency[9] and communication efficiency[10] have been theoretically estimated. The results obtained are comparable to those of similar amplitude quantized p.c.m. systems. The range of applications and the message grade expected in different systems may be estimated from the table and other figures given above.

## Conclusions

From the results and comparisons given above, it is seen that the proposed binary slope quantized p.c.m. system has improved characteristics over similar Δ-M and Δ-ΣM systems. The slope quantized system at 40kc/s p.r.f. has overall characteristics comparable to those of the currently used 6-digit amplitude quantized system. The power and communication efficiencies of slope quantized p.c.m. are higher than the corresponding amplitude quantized p.c.m. because, the coding characteristics being similar, the slope quantized system requires less bandwidth for transmission. Thus a medium grade circuit with an average signal-to-noise ratio of 25 to 30dB may easily be obtained with binary slope quantized p.c.m. using only a limited channel capacity of about 40 000 bits/sec. The overall circuits in slope quantized p.c.m. are very simple, many channels can be multiplexed, and wideband input signals, such as television and f.d.m. can be processed economically. The coder in slope quantized p.c.m. is simpler than those of Δ-M and Δ-ΣM, and even without companding the dynamic range is sufficient for processing of speech and f.m. signals (as used in f.m. telemetring). A further improvement in overall performance for the digital processing of signals may be obtained by using a ternary system, as has been reported elsewhere[11].


### Acknowledgment

The author is indebted to Prof. J. Das for his helpful guidance and encouragement. He also wishes to thank Prof. H. Rakshit, for his kind interest in the work and to Dr. S. N. N. Pandit for his helpful discussions.


## Appendix

To calculate the signal-to-noise characteristics of the system shown in Fig. 2, the frequency domain characteristics of quantities such as $f(t)$, $E(t)$, $O(t)$, $G_1(t)$, $B(t)$, $G_2(t)$, $f'(t)$ and $G_3(t)$ are taken as $f(\omega)$, $E(\omega)$, $O(\omega)$, $G_1(\omega)$, $B(\omega)$, $G_2(\omega)$, $f'(\omega)$ and $G_3(\omega)$ respectively. From the circuit equations, it is seen that:

$$E(\omega) = f(\omega) - O(\omega) \,.\, G_1(\omega)$$

or:

$$\begin{aligned} E(\omega) \,.\, G_3(\omega) &= f(\omega) \,.\, G_3(\omega) - O(\omega) \,.\, G_1(\omega) \,.\, G_3(\omega) \\ &\simeq f(\omega) - O(\omega) \,.\, G_1(\omega) \,.\, G_3(\omega) \\ &\simeq f(\omega) - f'(\omega) = N(\omega) \quad \ldots\ldots\ldots\ldots\ldots \quad (5) \end{aligned}$$

where $G_3(\omega) = 1$ within the signal frequency band and is equal to zero outside the band, and $G_1(\omega)$ and $G_2(\omega)$ are assumed to be similar transfer functions. For sinusoidal signal inputs, the error waveform is triangular in nature as shown in Fig. 1(d). The error amplitudes vary in a completely random fashion between the peaks of $\pm V_e$, where $V_e$ is the height of the output pulses $O(t)$, and its power density spectrum[12] is:

$$\begin{aligned} S_E(\omega) &= (\langle \nu^2 \rangle / \tau) \, |E_T(j\omega)|^2 \\ &= (\langle \nu^2 \rangle \tau)/4 \, [\sin(\omega\tau/4)/(\omega\tau/4)]^4 \quad \ldots\ldots \quad (6) \end{aligned}$$

where $\langle \nu^2 \rangle$ = mean square amplitude of the pulses;
$\tau = 1/f_r$

**TABLE 1**
**Comparison of different P.C.M. systems**

| System characteristics | | P.R.F. 40kc/s | | | 6-digit[1,4] A.Q.-P.C.M. | P.R.F. 60kc/s | | | 7-digit A.Q.-P.C.M. |
|---|---|---|---|---|---|---|---|---|---|
| | | Δ-M[2,4] | Δ-ΣM[3] | Binary S.Q.-P.C.M. | | Δ-M | Δ-ΣM | Binary S.Q.-P.C.M. | |
| Without compandor | Optimum signal-to-noise ratio for sinusoidal signal (dB) | 28 | 23 | 37 | 37 | 33 | 29 | 40 | 42 |
| | Dynamic range for minimum signal-to-noise ratio of 25dB (dB) | 10 | — | 17.5 | 19 | 13 | 9 | 20 | 29 |
| | Noise signal test (dB) | 20 | — | 27 | 28 | 24 | 20 | 30 | 33-34 |
| With compandor | Uniform signal-to-noise ratio (dB) | — | — | 27 | 27 | — | — | 30 | 30 |
| | Dynamic range (for the uniform signal-to-noise ratio (dB) | — | — | 32 | 33 | — | — | 34 | 37 |

$E_T(j\omega) =$ Fourier transform of a triangular pulse of unit height.

$= \tau/2\,[\sin(\omega\tau/4)/(\omega\tau/4)]^2$

Therefore, the noise power in the message band is (for $\sin x/x \simeq 1$ for $\omega \leqslant \omega_c$ and $\omega_c\tau \ll 1$) :

$$N^2 = 1/2\pi \int_{\omega_0}^{-\omega_0} S_E(\omega)\,.\,d\omega$$

$$= (\langle v^2\rangle/2)\,.\,(f_c/f_r)$$

and the r.m.s. noise voltage is:

$$N = (\langle v^2\rangle/2)^{1/2}\,.\,(f_c/f_r)^{1/2} \quad \ldots\ldots\ldots\ldots \quad (7)$$

The maximum signal built up by the binary 1/0 pulses, for a sinusoidal message signal frequency $f_m$, subject to the particular impulse response $r(t)$ shown in Fig. 1(a), may be estimated as:

$$[S]_{rms} \simeq \frac{kV_e}{8\sqrt{2}}\,.\,(f_r/f_m)\ ;\ k \leqslant 1 \quad \ldots\ldots \quad (8)$$

where $V_e$ is taken to be the same as the peak error signal. Therefore the system signal-to-noise ratio is:

$$[S/N] = \frac{kV_e}{8\sqrt{\langle v^2\rangle}}\,.\,\frac{f_r^{3/2}}{f_c^{1/2}\,.\,f_m} \quad \ldots\ldots\ldots \quad (9)$$

If now a rectangular probability distribution of the amplitudes of the error waveform is assumed, the mean square amplitude $\langle v^2\rangle$ is found to be $V_e^2/3$, and from equation (9):

$$[S/N] = 0{\cdot}22\ \frac{kf_r^{3/2}}{f_c^{1/2}\,.\,f_m} \quad \ldots\ldots\ldots\ldots \quad (10)$$

The equation is similar to that given by Jager[2] for Δ-M. But for minimization of errors a peak distribution approaching a truncated normal curve having the same r.m.s. value as the normal distribution, may be assumed. For such a truncated normal distribution, the mean square amplitude $\langle v^2\rangle$ is found to be $V_e^2/16$ and equation (10) is modified accordingly. The signal-to-noise normally obtained will be between the two limits given by the above two idealized distributions of error amplitudes.

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## Electronic Fog Warning Device for Motorways

Lancashire Dynamo Electronic Products (MI Group) has developed photo-electric equipment to detect and give warning of fog on motorways and other major roads. The cost of installing the equipment on a 10-mile stretch of motorway is about £1 500 to £2 000.

Staffordshire County Police and local authority officials saw the equipment demonstrated when fog was simulated using smoke generators.

The equipment consists of a rugged, weatherproof light source transmitter unit and a remotely mounted photocell receiver unit designed for operating over a distance of up to 1¼ miles.

An infra red beam is projected between the two units and if the beam is partially or completely cut off by fog a relay within the equipment will automatically be operated.

The equipment is connected direct to a fog warning sign which switches on automatically.

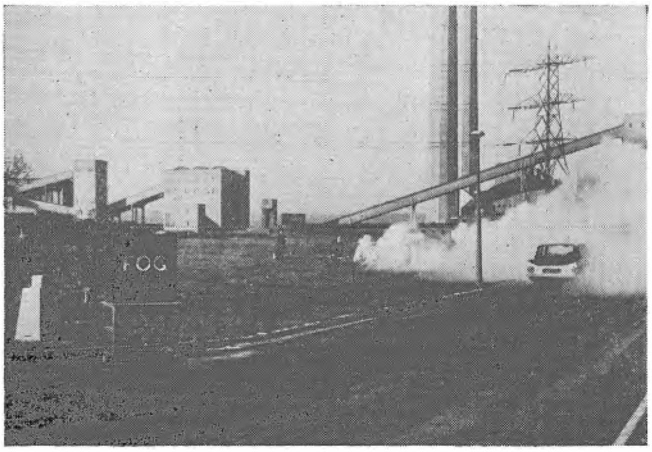
Alternatively, a signal can be transmitted through the roadside telephone system to a central control room where indicator lights on a control panel will light up showing the section of motorway which is fogbound. The control room operator can then switch on fog warning signs and inform police patrol cars. When visibility improves the warning device is automatically switched off.

The equipment uses modulated light to ensure that the photocell receiver unit only responds to the 'light' beam from the light source transmitter unit and is not affected by the varying ambient light, sunlight or motor vehicle headlights. It works equally well at night as during daylight.

A sensitivity control on the receiver unit allows simple adjustment of the desired operating level and the use of semiconductor components and 'fail to safe' features ensures a high degree of reliability.

The equipment was initially designed to detect fog and control foghorn warning systems for coastal navigation.

*Smoke generators being used to simulate fog for a demonstration of photo-electric fog detection equipment.*




## Computer with Micro-Integrated Circuits

A new range of computers known as System 4, intended for commercial, scientific and industrial applications, have been produced by English Electric-Leo-Marconi Computers Ltd using micro-integrated circuits.

The smallest computer in the System 4 range is the 4/10 which is designed for small commercial applications or for use as a satellite with larger computers.

It has a set of 26 instructions, a core store of from 4 096 to 16 384 bytes, and a cycle time, for 1 byte, of 1·5μsec (1 byte = 8 bits).

The 4/30 is the next size computer and has a core store of from 16 384 to 65 536 bytes with a cycle time for two bytes of 1·5μsec. It has a set of 41 instructions, FORTRAN and COBOL compilers and peripheral simultaneity—17 operations including the central processor can be carried out at the same time.

The 4/50 is a medium sized computer with a standard set of 100 instructions, supplemented by 44 floating point instructions enabling the most complex mathematical and scientific problems to be handled. Other facilities include a core store of from 16 384 to 262 144 bytes, with a cycle time of 1·4 microseconds for 2 bytes, a very fast 120 nanosecond "scratch pad" store, CLEO, COBOL, FORTRAN and ALGOL compilers, and 15-way multiprogramming.

The 4/70 is the largest computer in System 4. It can handle the most demanding commercial and scientific applications. A core store of exceptional capacity (65 536-1 048 567 bytes) has a cycle time for 4 bytes of 1 microsecond, interleaved to give an effective cycle time of 0·7 microseconds.

# An Operational Transistor Amplifier without Automatic Drift Correction

By M. Carapic* and D. Velasevic*

*This article deals with the design and analysis of an operational transistor amplifier which does not use automatic drift correction.*

*A low order of drift is ensured by using matched pairs of bipolar transistors in two differential stages. The drift of the amplifier, used as an invertor, is 150µV for a temperature range of 30°C, and the bandwidth 600kc/s.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

THE drift of d.c. amplifiers may be decreased by different techniques such as: compensation by thermistors or diodes, differential amplifiers, direct conversion of d.c. to a.c. voltages, and correction by chopper amplifiers. The simplest stabilization can be realized by non-linear elements, but it is not practical and efficient. So far, transistor differential amplifiers do not possess adequate stability[1], and direct conversion does not offer a satisfactory frequency response[2]. Drift correction appeared to be the most reliable method for the stabilization. It was introduced by Goldberg[3], and later modified by Blecher[4] to be used for transistors. Slaughter[5] and Okada[6] have analysed it in detail. When a transistor chopper is used for the convertor, the drift stability of the amplifier[7] is about 7µV/°C.

Recent progress in transistor technology of matched modular units makes possible reducing the equivalent input drift close to the drift of the correction element (transistor chopper). In this way, the correction amplifier can be eliminated and the construction greatly simplified.

### Basic Conception of the Amplifier

Fig. 1 shows a block diagram of the operational amplifier.

The first amplifier $A_1$ is to provide low drift, sufficient amplification, and the required gain-frequency characteristic (i.e., a narrow pass-band) which is achieved by the capacitor $C$ and high output impedance $R_o$. The amplifier $A_2$ due to its high amplification and wide pass-band provides the accuracy of the overall amplifier.

The transfer function of the amplifier is as follows:

$$e_o/e_i = -(Z_f/Z_i)\,\frac{1}{1+\dfrac{Z_fR_{in}+R_{in}Z_i+Z_iZ_f}{A_vR_{in}Z_i}} \quad \ldots\ldots (1)$$

where $A_v$, the open loop voltage amplification, is given by:

$$A_v \simeq \left(1+\frac{A_1}{R_oCp+1}\right).A_2 \quad \ldots\ldots\ldots\ldots (2)$$

The transistor stages in the common emitter configuration have the constant product of the amplification and input impedance, i.e., $A_vR_{in}$ = Const. From expression (1), for the given values of the impedances $Z_i$ and $Z_f$, it can be seen that the input impedance is to be lower. However, regarding the stability, a system with high input impedance is better[8], especially in applications requiring a change of the operational amplification $A = -(Z_f/Z_i)$.

For this reason, it has been decided to take into consideration an amplifier with high input impedance. The complete circuit diagram of the operational amplifier is shown in Fig. 2.

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### Analysis of Stability

D.C. amplification of the open loop amplifier is about 150dB. Such a high amplification makes the stability problem more difficult. Therefore, the frequency and phase characteristics must be carefully shaped. As is known, high

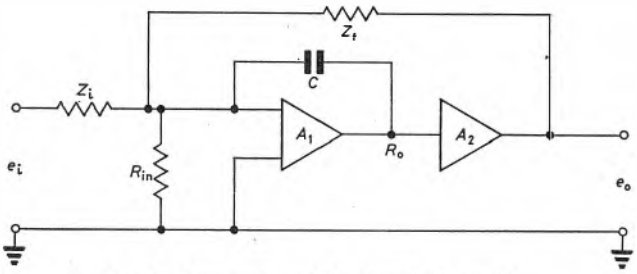



Fig. 1. Arrangement of the operational amplifier

frequency transistor characteristics are mainly determined by the output capacitance $C_o$ and cut-off frequency $f_\alpha$. While the capacitance $C_o$ is fairly constant for all transistors of the same type, the frequency $f_\alpha$ is changed with temperature and working conditions in the circuits. All transistors used in the amplifier have $f_\alpha > 20$Mc/s which makes it possible to consider the amplification $\alpha$ independent of frequency. In this way the number of variables participating in frequency shaping is decreased. Then, the existing $RC$ networks are local feedback circuits consisting of capacitances $C_o$ and output impedances of the preceding stages and interstage couplings.

Fig. 2 shows that three compensation circuits are incorporated in the amplifier $A_2$. The capacitances $C_o$ of the stages with the bases on low impedances are not taken into account.

The phase margin, as it is shown in Fig. 3, is 70° making the amplifier very stable. The gain margin is also high, although it is not shown in the diagram. The bandwidth of the amplifier, as an invertor, is 600kc/s.

The gain-frequency characteristic of the loop gain of the integrator is also shown in Fig. 3, the input resistance being $R_i = 10^4\Omega$ and feedback capacitance $C_f = 0{\cdot}1\mu F$. At low frequencies the loop gain is above 70dB ensuring high accuracy of the integrator. At high frequencies the characteristic is shifted 6dB to the characteristic of the invertor.

The phase characteristic of the integrator at low frequencies is shifted for 90° from the corresponding

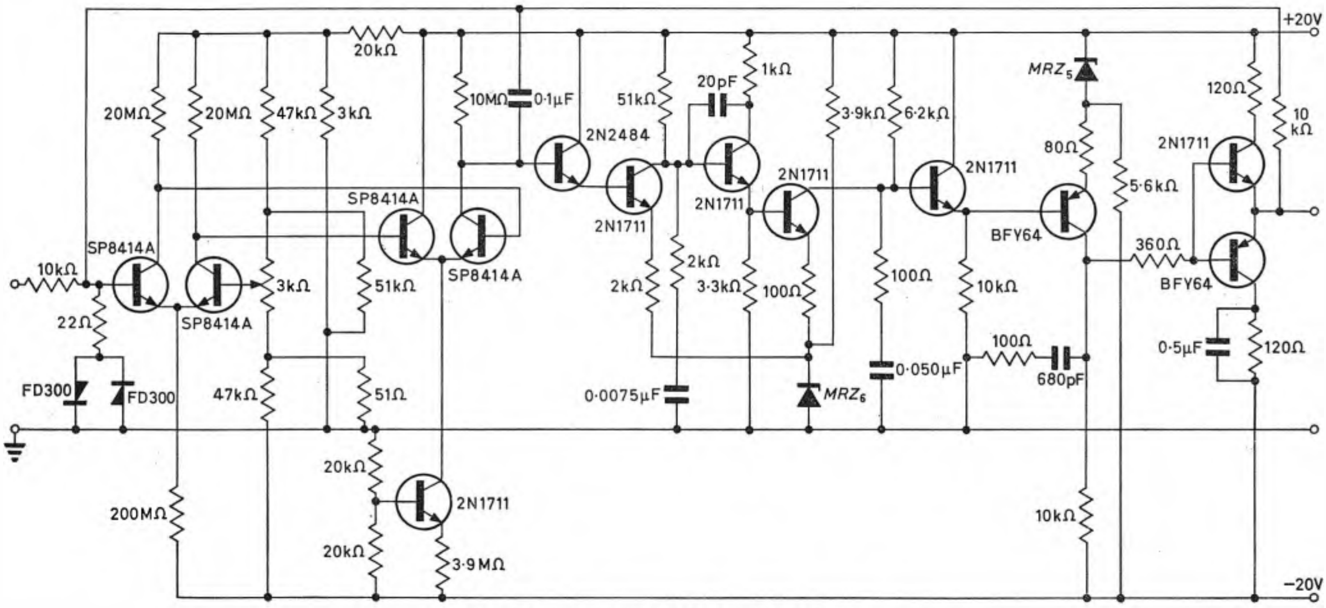



**Fig. 2. Complete circuit of operational amplifier**

characteristic of the invertor, but at high frequencies they coincide. The phase margin is somewhat decreased but still enough to provide good stability. The fact that the same compensation networks are used in the invertor and integrator shows, the advantages of the amplifier conception.

**Temperature Stability**

The first stage of the amplifier is dominant regarding zero drift. Its differential amplification is only forty, but the next stage, being differential, is also very stable; therefore, its influence on the drift can be neglected. The drift of the operational amplifier is determined by equivalent voltage and current generators at the input of the first differential stage, as Konigsberg[6] has shown for a single ended amplifier (see Fig. 4).

The output drift $E_{OM}$ represents the maximum effects of all misalignment sources. If the base to emitter voltages and inverse saturation currents are considered to be dominant, the voltage and current offsets of the first differential stage will be respectively:

$$E_M \simeq \Delta V_{be} + (R_L/A_d) \,.\, \Delta I_{co}, \qquad I_M \simeq \Delta I_{co} \qquad \ldots\ldots\ldots\ldots (3)$$

where $R_L$ is the collector load, and $\Delta$ denotes total differential temperature changes.

Due to these misalignment sources the output drift is as follows:

$$E_{OM} = E_M\,(1 + (Z_f/Z_i)+(Z_f/Z_{in})) + I_M Z_f \quad \ldots\ldots (4)$$

If the expression (4) is to be analysed, it is necessary to separate all influences and to confirm the conclusions by measurement. The last term $I_M Z_f$ can be neglected if the feedback resistance is low, i.e., $Z_f \leqslant 100\text{k}\Omega$. The input impedance being very high, $Z_{in} \gg Z_f$, expression (4), for $Z_f = Z_i$, $A_D = 40$, and $R_L = 20\text{M}\Omega$, yields:

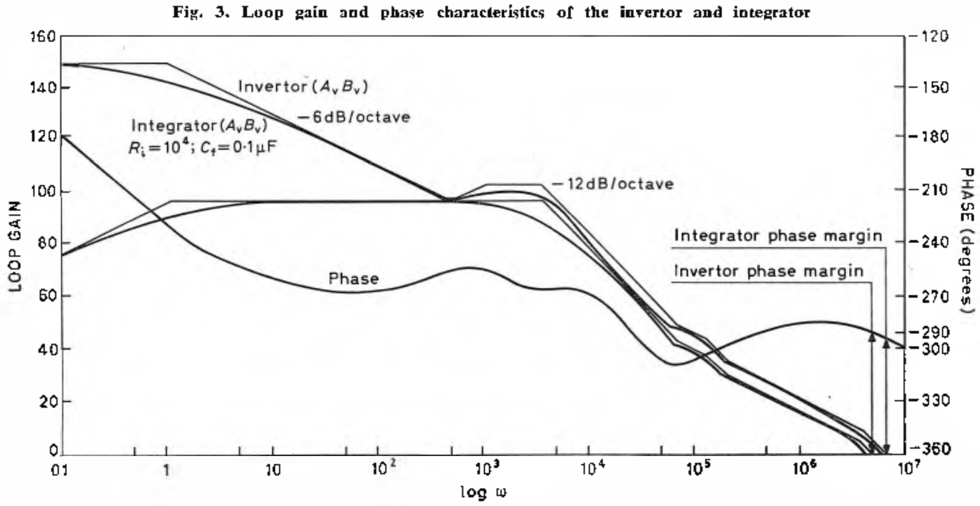
$$E_{OM} = 2\,(\Delta V_{be} + 5\,.\,10^5\,.\,\Delta I_{co}) \quad \ldots\ldots\ldots (5)$$

The output drift of $5\mu$V/°C was found by measurement in ten operational amplifiers. Typical offsets $\Delta V_{be}$ of dual transistors (Fairchild SP 8414A) are $1\mu$V/°C and, consequently, $\Delta I_{co} = 3\,.\,10^{-12}$A/°C.

On the other hand, the output drift of the integrator depends on the expression $I_M Z_f$, i.e.:

$$E_{OM} = \left[\,2{\cdot}5\,.\,10^{-6}\,(1 + (t/C_iR_i)) + \frac{3\,.\,10^{-13}t}{C}\,\right] (\text{V}/^\circ\text{C/sec}) \quad \ldots\ldots\ldots\ldots (6)$$

**Fig. 3. Loop gain and phase characteristics of the invertor and integrator**




Now, operational amplifiers using automatic drift correction by transistor choppers could be compared with amplifiers using dual transistors only. Transistor chopper zero offset changes with temperature. These changes cannot be compensated by an a.c. amplifier. Silicon transistors which are preferable to germanium for transistor choppers yield $7\mu$V/°C at the output of the adder[7] which is comparable with the drift of the amplifier proposed. Integrators using automatic drift

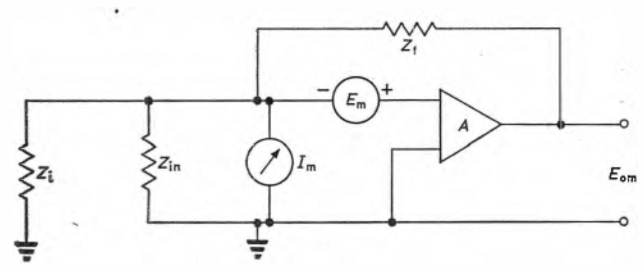



**Fig. 4. Equivalent amplifier representation showing misalignment generators $E_M$ and $I_M$**

correction are not sensitive to $I_{co}$ when a capacitor of high quality is at the input of the amplifier. However, for good frequency response the capacitor needs to be of high value and the leakage still represents a big problem. Therefore, it is difficult to say if there is any advantages in an integrator with no drift correction. From the expression (6) it can be concluded that for all practical values of calculating elements the drift is low. Experimentally it has been confirmed that the drift is not higher than $100\mu V/sec$ at normal ambient temperatures for $R_i = 100k\Omega$ and $C_f = 0{\cdot}1\mu F$; this is also comparable with the drift of an integrator using automatic drift correction.

### Other Characteristics of the Amplifier

The high input impedance of the amplifier is due to very low collector currents of the first stage. It can be shown that it is:

$$Z_{in} = 2r_e(\beta + 1) \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

where $r_e$ is the intrinsic emitter resistance and $\beta$ the current amplification. The resistance $r_e$ is very high at low currents and for the first stage which is working at $5 . 10^{-8}$A it is $500k\Omega$. The current amplification of the transistors is still high so that an input impedance of $Z_{in} = 100M\Omega$ was obtained.

The time response was experimentally determined and is $5\mu$sec. The overshoot is 15 per cent.

The noise of the amplifier is about $70\mu V$.

### Conclusion

The basic characteristics of the operational amplifier with dual transistors but without drift correction show that the amplifier is of a high quality. By eliminating the correction amplifier, zero drift is not enlarged and spikes generated by the chopper do not exist at all. A phase margin, i.e., high frequency stability for a wide pass-band, is achieved by careful frequency shaping.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# A Simplified Approach to the Design of Transistor Multivibrator Circuits


By P. Mars, B.Sc.



*A procedure is described for the design of transistor multivibrator circuits having specific values of pulse duration and amplitude, recovery time of the initial state and relaxation time temperature stability. Since the reactive components of the transistor input impedance are neglected the design procedure is only applicable to multivibrator circuits with relatively large relaxation times. Conducting transistors are assumed to be in saturation.*



(Voir page 65 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)


THREE basic forms of transistor multivibrator circuits are shown in Figs. 1, 2 and 3. Fig. 1 is the conventional symmetrical collector-coupled astable transistor multivibrator circuit using pnp transistors. Fig. 2 is a normal one-shot monostable circuit and Fig. 3 is the circuit for an emitter-coupled Schmitt trigger circuit.

Considering Fig. 1 when transistor $VT_1$ switches from 'off' to 'on', its collector voltage has a positive step from $-V_{cc}$ to $V_{ce(sat)}$. This positive step is transferred to the base of transistor $VT_2$. The time for the base-emitter voltage of transistor $VT_2$ to relax into the active region is given by the expression

$$t_2 = R_2C_2\ln 2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

This assumes that the input resistance of the cut-off transistor $R_r$ is much greater than $R_2$.

### Relaxation Time Temperature Stability

The principle effect producing temperature instability of the relaxation time in multivibrators is the amount of shunting on the *RC* coupling components by the input of the quasi-stable state transistor. If $R_r$ is not much greater than $R_2$ then the circuit time-constant is the product of $C_2$ and the parallel combination of $R_2$ and $R_r$. The initial value of the exponential is $V_{cc}R_r/R_r + R_2$. Solving for the time required for the reverse voltage exponential to relax to zero volts gives

$$t_2 = C_2 . \frac{R_2 . R_r}{R_2 + R_r} . \ln \frac{R_2 + 2R_r}{R_2 + R_r} \quad \ldots\ldots\ldots\ldots (2)$$

In saturated multivibrator circuits the relaxation time decreases monotonically with increasing temperature. Hence the variation of the relaxation time over a specified temperature range may be represented by the expression

$$F_T = \frac{t_2' - t_2^*}{t_2'} \quad \ldots\ldots\ldots\ldots\ldots\ldots (3)$$

The notation used here is that the value of a parameter at the minimum temperature limit is indicated by a

* *City of Liverpool College of Technology.*

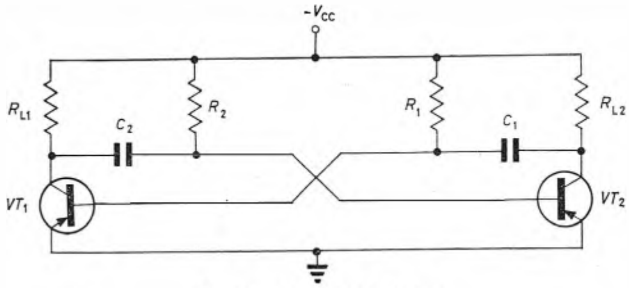



Fig. 1. Astable multivibrator

prime ($'$), and the value of the parameter at the maximum temperature limit by an asterisk ($*$).

Differentiating equation (2) and using finite increments gives from equation (3)

$$F_T = \frac{R_2 \,.\, (R_r' - R_r^*)}{(R_r^* + R_2)\,(2R_r' + R_2)\ln\left\{\dfrac{R_2 + 2R_r'}{R_2 + R_r'}\right\}} \quad \ldots\ldots (4)$$

Under the assumption that $R_r' \gg R_r^*$

and $R_r' \gg R_2$

Equation (4) reduces to

$$F_T \simeq \frac{R_2}{(R_r^* + R_2)\,2\,.\ln 2} \quad \ldots\ldots\ldots\ldots\ldots (5)$$

**Recovery Time**

An important factor in the consideration of non-linear circuit functions is the problem of designing a circuit to return, or recover to its original or desired state. The circuit must recover in order that the particular non-linear response can be consistently repeated within prescribed time intervals. A recovery time factor $F_R$ will be introduced given by

$$F_R = (t_1/t_2) \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

and the recovery time $t_1$ will be defined as four times the product of the time-constant associated with the discharge of the time assigning capacitor.

$$\text{i.e. } t_1 = 4R_{L2}C_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

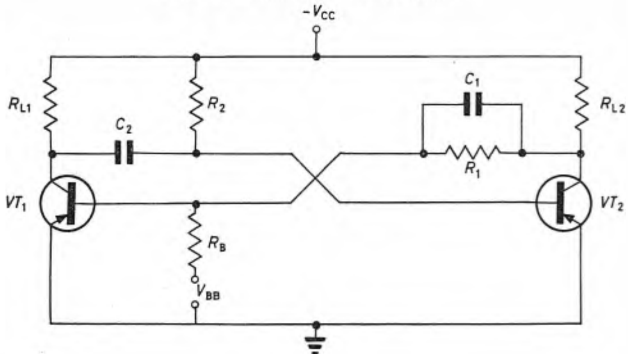
For the Schmitt trigger circuit of Fig. 3.
recovery time $t_1 = 4(R_{L2}+R_E)\,.\,C_2$

Usually however, $R_{L2} \gg R_E$ and hence it will be assumed that equation (7) holds for both the circuits of Fig. 2 and Fig. 3. In the astable circuit of Fig. 1 $t_1$ will represent the charging time of either capacitor $C_1$ or $C_2$.

Simultaneous solution of equations (1), (5), (6) and (7) produces the relationship

$$R_{L2} \simeq \frac{F_T \,.\, F_R \,.\, R_r^*(\ln 2)^2}{2(1-2\,.\ln 2\,.\,F_T)} \quad \ldots\ldots\ldots\ldots\ldots (8)$$

Fig. 2. Monostable multivibrator




Equation (8) is essentially a first draft design formula which simplifies the design and makes it more accurate.

**Design Procedure**

For the desired output pulse amplitude a supply voltage $V_{cc}$ is chosen and a transistor $VT_2$ selected with a known value of $R_r^*$. For given values of $F_R$ and $F_T$ substitution into equation (8) gives the value of $R_{L2}$ required. If the calculated value of $R_{L2}$ is satisfactory the resistance $R_2$ and capacitance $C_2$ required to produce the specified values of $t_1$ and $t_2$ may be evaluated from equations (1) and (7).

**Typical Design**

Consider the design of a monostable multivibrator Fig. 2 having the following specifications.

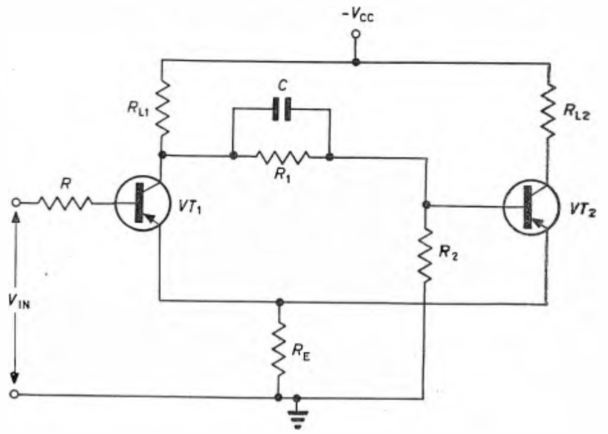



Fig. 3. Emitter-coupled trigger circuit (Schmitt trigger)

$V_{cc} = -10$V, $t_2 = 1$msec, $F_T = 0{\cdot}1$ and $F_R = 0{\cdot}1$ over a temperature range $-55$°C to $+55$°C.

Suppose the transistors used have an $R_r \geqslant 0{\cdot}5$MΩ at the upper temperature limit.

Equation (8) gives $R_{L2} \simeq 1{\cdot}4$kΩ.

If the corresponding collector current

$$I_c = (V_{cc}/R_{L2}) \simeq 7{\cdot}2\text{mA}$$

is permissible, equations (7) and (1) give

$$R_2 \simeq 80\text{k}\Omega \text{ and } C_2 \simeq 0{\cdot}018\mu\text{F}.$$

**Conclusions**

An equation has been derived relating the parameters usually specified in the design of multivibrator circuits. Considerations required in the calculation of transistor quiescent conditions have not been dealt with but may be found elsewhere[1,2]. It should be noted that the choice of $F_T$ and $F_R$ is essentially a compromise. In the case of the astable multivibrator shown in Fig. 1 the factor $F_T$ could be reduced by decreasing the values of $R_1$ and $R_2$. However this step would also reduce the oscillation period unless the values of $C_1$ or $C_2$ were increased. The effect of increasing the values of the coupling capacitors would be to increase the recovery time $t_1$, although this could be compensated by a reduction in $R_{L1}$ and $R_{L2}$.

**Acknowledgments**

Thanks are gratefully acknowledged to the Head of the Department of Electrical Engineering of the Liverpool College of Technology for permission to publish this article.

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# Noise in the Metal-Oxide-Semiconductor Transistor

By S. M. Bozic*, Dipl. Ing.

*The experimental results for the noise figure of the m.o.s. transistor are given in the frequency range 30 to 1 600kc/s and the source resistance 600Ω to 100kΩ. The effect of the operating point on the noise figure is also shown. The conclusions reached are briefly discussed.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

THE metal-oxide-semiconductor (m.o.s.) transistor, a modified version of the field-effect transistor (f.e.t.), is suitable for a wide range of applications[1]. The agreement between the noise theory and the experiment for the f.e.t. has been reported to be good[2]. However, it has also been found that low frequency noise well in excess of the theoretical level is present having a $1/f$ type spectrum[3].

There is not much information published about the noise properties of m.o.s. transistors. The noise figure of this device, according to some manufacturers, can be as low as a few decibels but for rather large values of source resistance (1MΩ) and at low frequencies (1 to 10kc/s). On the other hand there are opinions that they are not, at present low noise devices[4]. In order to find more information about the noise in m.o.s. transistors, their noise figure has been measured under different conditions. The measuring arrangement and results of this work are described below.

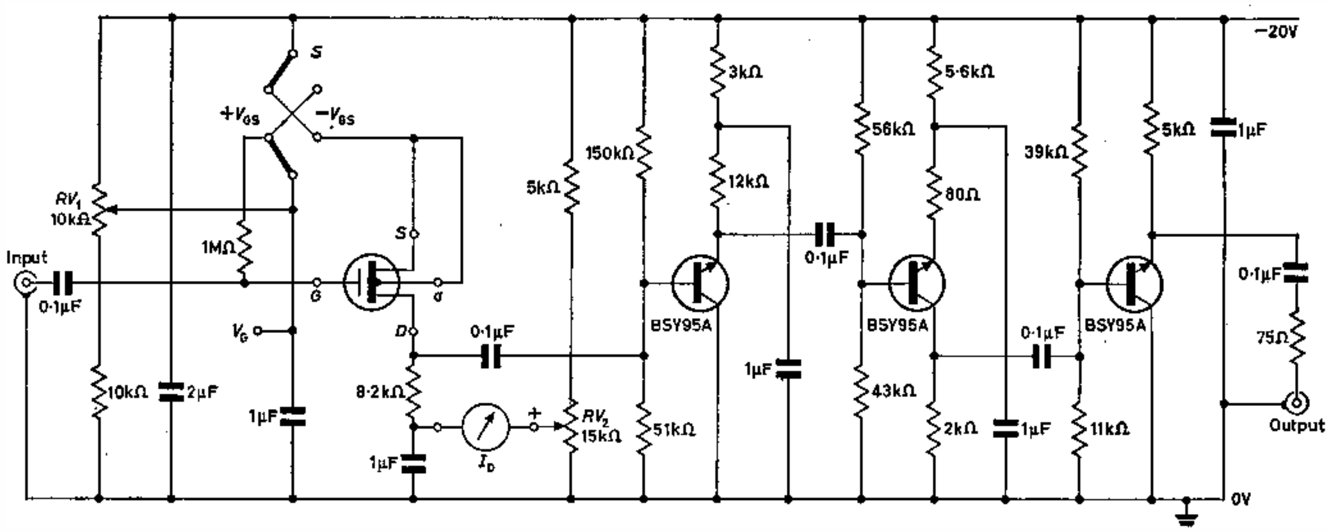


Fig. 1. The pre-amplifier used for the noise measurements

## Measuring Arrangement

The pre-amplifier used for the noise figure measurements is shown in Fig. 1. The first stage is the m.o.s. transistor under test. The gate-source and drain-source voltages can be set by means of the potentiometers $RV_1$ and $RV_2$ respectively. The switch $S$ serves the purpose of changing the polarity of the gate-source voltage. The rest of the pre-amplifier contains three transistor stages which provide additional gain and convenient impedance levels. The output from the pre-amplifier was fed into an 'Airmec' waveform analyser, type 853. The noise figure was measured by the standard method[5] using a CV2398 for the noise diode.

## Experimental Results

The noise figure ($F$) has been measured for several samples of m.o.s. transistor type 95BFY (Mullard) and one depletion type sample of DPT201 (Pacific Semiconductors). The operating conditions in each case are given in the graphs. All the measurements were performed in common source configuration with the substrate connected to the source.

The results for the samples of 95BFY are shown in Figs. 2 and 3 for the DPT201 in Fig. 4. They represent $F(f, R_s)$ measured in the frequency range 30kc/s up to 1·6Mc/s and the source resistance 600Ω to 100kΩ.

The common feature of these results is the minimum which decreases and shifts towards the lower frequencies as the source resistance increases. The rising slope towards the lower frequencies is approximately 3dB/octave which corresponds to a $1/f$ law. The slope towards the high frequencies is higher than 3dB/octave. The results, in Fig. 3, for three samples of type 95BFY operated from $R_s = 5{\cdot}1\text{k}\Omega$ show considerable spread in the noise figure levels and their minima.

Further investigations have been carried out to determine the noise figure behaviour as a function of the operating d.c. conditions. One set of the results $F(I_d, f)$ is presented in Fig. 5. The fixed frequencies for 95BFY have been chosen at, below and above the frequency of the minimum noise figure. The results show that the noise

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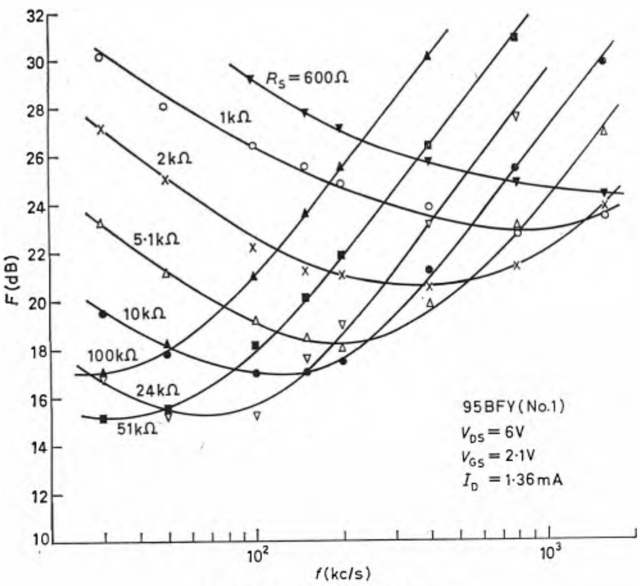



**Fig. 2. Results for 95BFY**

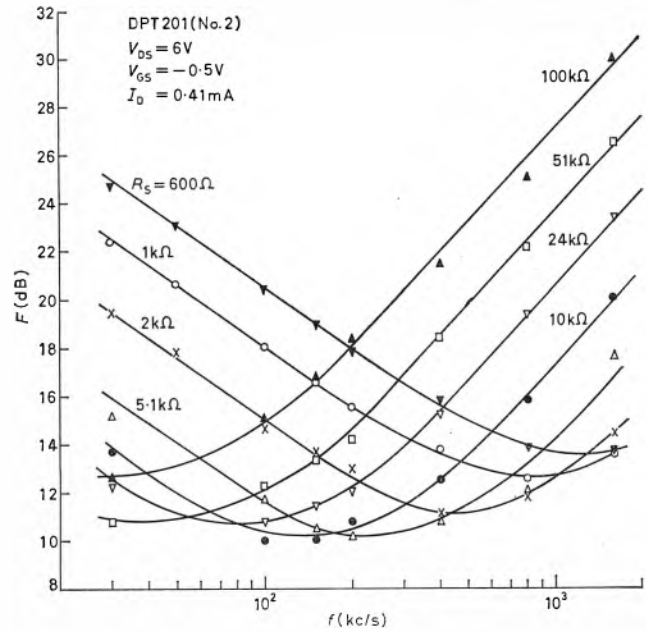



**Fig. 4. Results for DPT201**

figure has wide minima with respect to the drain current and that the shapes of the curves at the three frequencies (30, 200 and 1 600kc/s) do not differ appreciably. The sample of DPT201 shows lower noise than the 95BFY but the general behaviour is similar. The measurements have also shown that the noise figure is independent of the drain voltage ($V_d$) for the DPT201. The same has been found for 95BFY for $V_d > 10V$ but the noise figure increased approximately 2dB when this voltage was decreased from 10V to 2V.

In order to make a definite statement on the noise properties of these two types of m.o.s. transistors one would have to examine a larger number of samples since there is, as Fig. 3 shows, a significant spread between the samples of the same type.

The conclusion is that in order to achieve a low noise figure with the present types of m.o.s. transistors they have to be operated from a large source resistance ($> 100k\Omega$) and at low frequencies ($<10kc/s$). It seems, however, that a hybrid connexion such as the cascode could avoid the

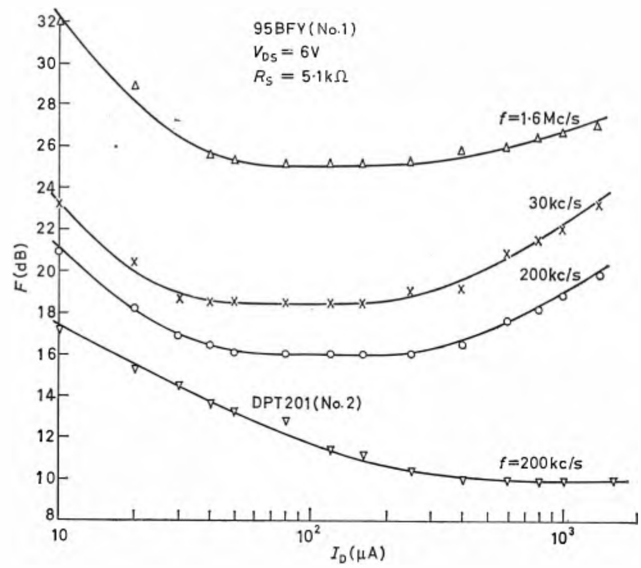
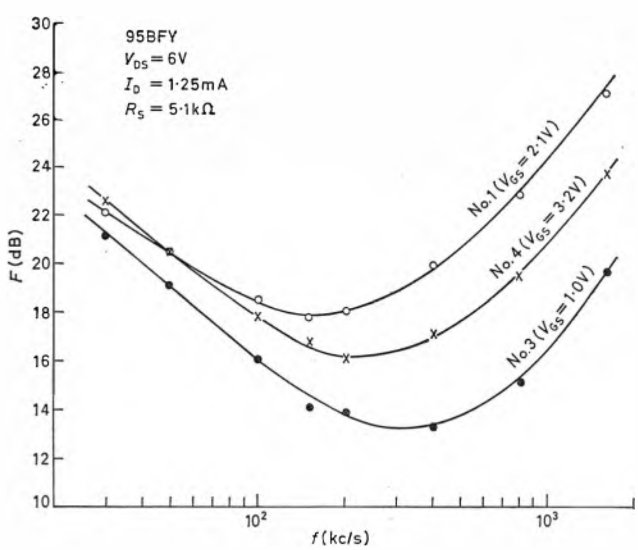



**Fig. 5. Noise figure as a function of d.c. conditions**

**Fig. 3. Results for three samples of 95BFY with $R_s$=5·1kΩ**




above limitations. In this case the first stage would be a diffusion (or drift) type transistor operating in the common emitter configuration followed by an m.o.s. transistor in the common gate connexion.

**Acknowledgments**

The author wishes to thank the General Electric Co. Ltd for the permission to publish the results of this investigation. Also thanks are due to Mr. R. Morrall for help in the measurements.

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# Crystal Oscillator Using Integrated Circuit Amplifiers

By A. H. James*

*The article describes an oscillator which uses a quartz crystal as a series resonant element in conjunction with solid state amplifiers. A compact and simple circuit results, and has a frequency stability of 60 parts per million between $-50°C$ and $+100°C$, and 0·5 parts per million per volt of h.t. change.*

(Voir page 63 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 70)

THE oscillator described here was produced to meet a situation where size and weight were of maximum importance, and where careful matching of temperature coefficients would have been necessary if an *LC* oscillator had been used, because of the very wide temperature range involved.

The simplest solution is to use a quartz crystal as the frequency determining component, and silicon integrated circuit units in the maintaining amplifiers.

For some time solid state units, containing a complete circuit in transistor type can, have been available for use in logic circuits, and more recently a range of units for use in linear circuits has been introduced. In addition to being very much smaller physically than conventional circuits, these units offer possibilities of improved reliability and, eventually, reduced cost. The methods by which they are manufactured require that large numbers of each type of circuit must be produced before costs can be reduced. Logic circuits lend themselves well to this approach since, normally, large numbers of a few types of circuit are needed. Linear circuits are more usually designed on an individual basis, but they can be a practical proposition in integrated circuit form if selection can be made from a small range of multipurpose units. These must be designed so that, by rearrangement of can connexions and possibly with the addition of one or two external components, a number of different circuit functions can be performed.

The circuit described here is an example of this approach. Although the Ferranti ZLA10 unit (Fig. 1) is basically an amplifier, it is used here

(a) As an oscillator, when a crystal, a resistor and a diode are added as external components, together with capacitors for coupling and decoupling.

(b) As a buffer stage, with re-arranged connexions (Fig. 2).

The oscillator circuit (Fig. 3) consists basically of a two stage transistor amplifier with limiting, the crystal being included in a positive feedback loop. It is based upon a design by P. J. Baxandall[1] for obtaining a performance of very high order from an inexpensive crystal.

* *Royal Radar Establishment.*

## Circuit Details

Quiescent d.c. conditions in the oscillator are determined by negative feedback from $VT_2$ emitter to $VT_1$ base. The base-emitter voltage of a silicon transistor when conducting is approximately 0·7V, this potential being used to set up the standing currents indicated in Fig. 3.

The drive from $VT_2$ collector to $VT_1$ base via the crystal provides positive feedback, and determines the frequency of oscillation. Current swing in $VT_2$ is from 3·5mA to 14mA, and the voltage swing at $VT_2$ collector is limited to about 1·2V peak-to-peak, by the action of the two diodes. This limiting action is most important, and acts as

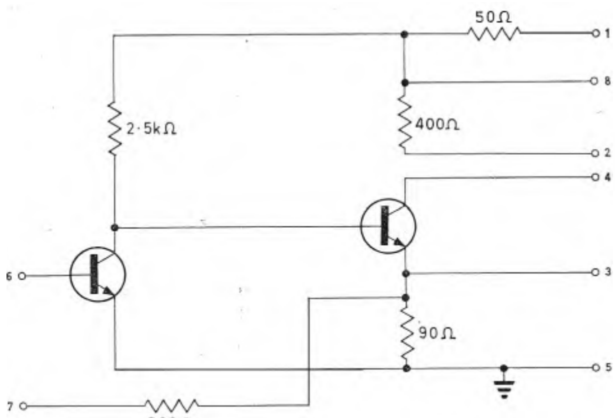
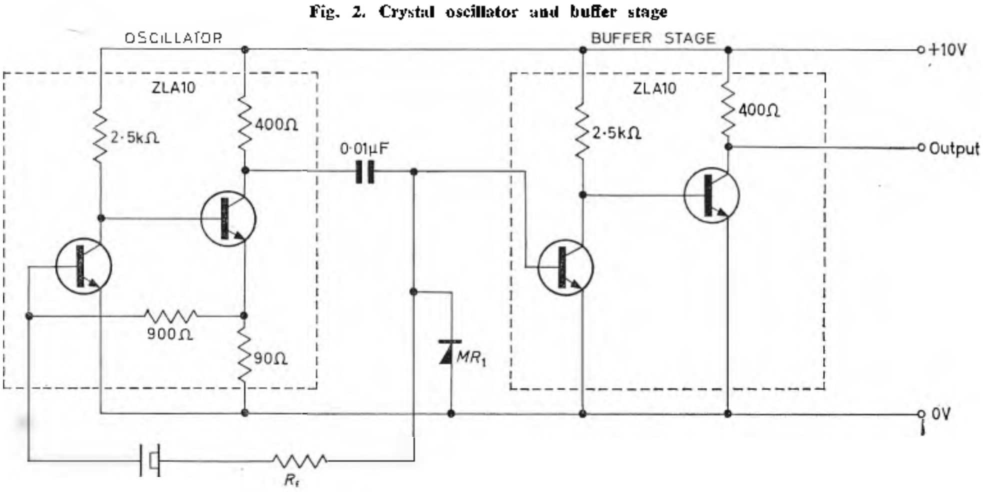



**Fig. 1. Ferranti Microlin amplifier ZLA10**

**Fig. 2. Crystal oscillator and buffer stage**



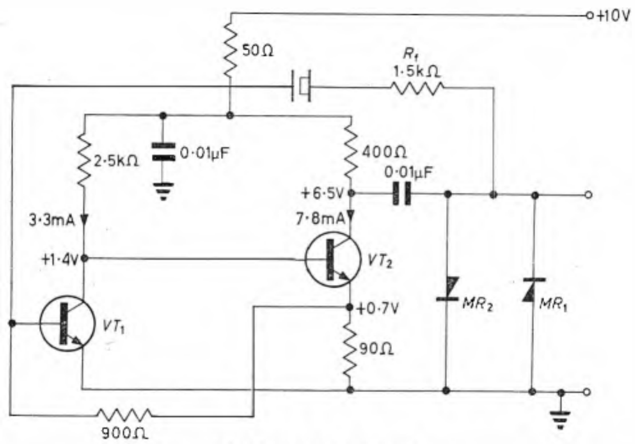



Fig. 3. Basic oscillator

a form of a.g.c., enabling the amplifying unit to work in a linear manner, and giving a worthwhile improvement in frequency stability. The resistor in series with the crystal ($R_f$) determines the amount of positive feedback, and also reduces dissipation in the crystal. The value is not critical; its upper limit, about 3kΩ, is that which causes oscillation to cease, while the lower limit, about 700Ω, is determined by overloading of the amplifier, which can be observed at $VT_2$ emitter. Dissipation in a crystal has a marked effect upon long term stability, and from this point of view a low value of feedback current is desirable.

### Output Stage

Sufficient drive is required from the oscillator to feed a logic unit from the Ferranti 'Micronor' series, and to achieve this a second ZLA10 unit is used as a buffer stage (Fig. 2). In this arrangement one diode is dispensed with, the base-emitter diode of $VT_3$ being used for limiting. With a logic element as load for this buffer stage, the oscillator produces a square wave with edge times of 7nsec and 12nsec respectively as $VT_4$ is turned on and off.

### Performance

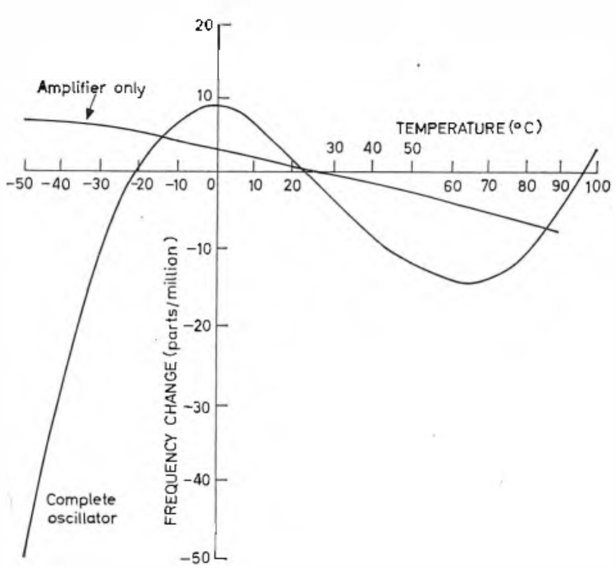
The oscillator has operated successfully at frequencies between 1Mc/s and 14Mc/s using crystals in their fundamental mode. The figures for stability with temperature and h.t. were obtained using a 5·5Mc/s, AT cut crystal.

To provide a simple and convenient reference two similar oscillators were constructed. One was maintained at constant temperature and fed from stabilized supplies, while conditions on the second oscillator were varied; temperature between −50°C and +100°C, voltage ±20 per cent.

Fig. 4 shows the variation of frequency with temperature. The temperature at 25°C is taken as reference and the frequency change is less than ±15 parts per million between −30°C and +100°C. This change however, increases very rapidly at lower temperatures and reaches 50 parts per million at −50°C. Although both crystal and amplifier contribute to it, the dominating factor is the crystal temperature characteristic.

When stability with h.t. voltage is considered the situation is reversed, and the frequency shift is caused by changing conditions in the amplifiers. A very marked improvement in performance is shown if limiting diodes are employed and overloading is avoided. (Fig. 5).

Fig. 4. Change of frequency with temperature



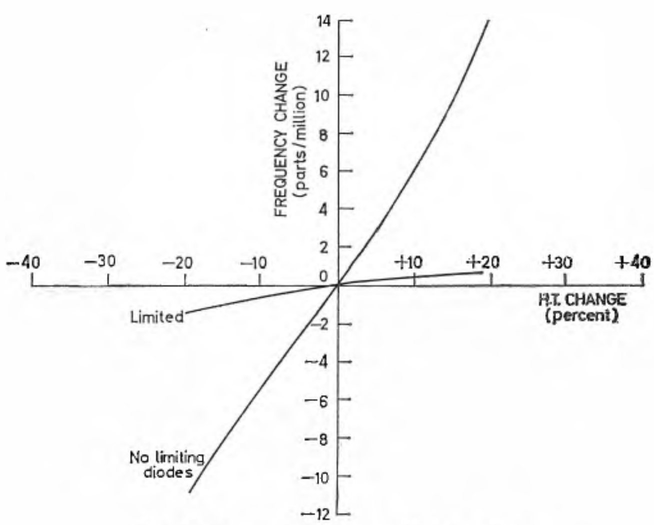




Fig. 5. Change of frequency with h.t. voltage

### Conclusions

The frequency stability of the oscillator is much better than that which could be achieved by an oscillator using inductors and capacitors, because in the latter case it would be necessary to attempt to match the temperature coefficients of inductors and capacitors over a wide temperature range.

A superior performance could be obtained from a crystal oscillator using conventional techniques. This is in no way due to imperfections in the solid state units, but to the circuit simplifications introduced to enable a simple and compact oscillator to be made.


#### Acknowledgments

The writer wishes to acknowledge with thanks the help and advice given by Mr. P. J. Baxandall and Mr. R. C. Bowes while he was carrying out this work.

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# LETTERS TO THE EDITOR

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

### Triggered 'Energy Reclaim' Inductive Time-Base

DEAR SIR,—It sometimes occurs that an inductive time-base designed for isochronous operation on the 'energy reclaim' principle is required to be triggered by a sequence of pulses whose repetition intervals irregularly depart from the mean by small amounts. It would nevertheless be unacceptable for

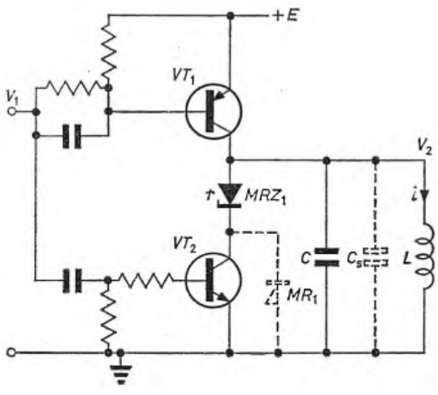



**Fig. 1. Adaptation of conventional circuit**

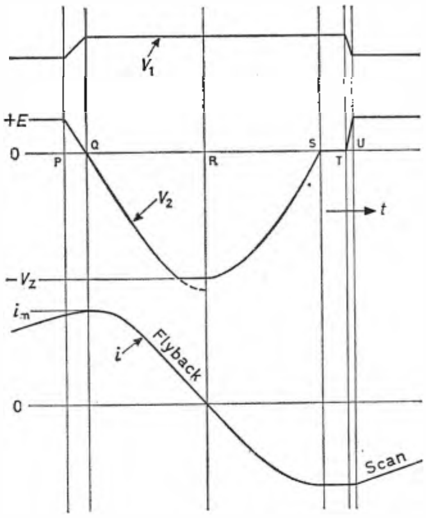



**Fig. 2. Waveforms of Fig. 1**

the c.r.t. spot to start from a different co-ordinate on the screen for each different interval between pulses.

An adaptation of the conventional circuit so as to satisfy this situation is shown in Fig. 1, and the corresponding waveforms in Fig. 2. During the period of scan, the required constant voltage is impressed across inductor, $L$, (representing the deflexion coils, and transformers, if any) by the regular supply switch, $VT_1$, held 'on' by the more negative part of waveform $V_1$. If the next positive (or flyback) part is late, the current in $L$ will have risen to above the value appropriate to the mean rate. Thus the stored inductive energy, $\frac{1}{2}Li_m^2$, will also be high and, after the half cycle oscillation of flyback, would reappear as current $-i_m$ in $L$, giving excessive deflexion at the start of the next scan.

$V_1$ is applied via a speed-up $RC$ circuit to switch $VT_1$ off quickly at time $P$, but complementary transistor $VT_2$ on more gradually as the result of its limited rate of rise. $VT_2$ does not conduct until after time, $Q$, at which $V_2$ goes negative. The circuit made up by $L$ in parallel with $C$ and strays $C_s$ is no longer loaded, and $(C + C_s)$ start a quarter of a sinusoidal oscillation to take over the energy previously stored in $L$. The resulting voltage swing takes $V_2$ considerably negative to earth, sufficiently for Zener diode (or diodes in series), $MRZ_1$, to conduct via $VT_2$ bottomed and inverted (or via optional diode $MR_2$ shown dotted). At time $R$, $V_2$ has thus been limited to $-V_z$, the running voltage of $MRZ_1$ reverse conducting, and the electrostatic energy therefore to $\frac{1}{2}(C + C_s)V_z^2$, a standard value a little lower than the lowest expected value of $\frac{1}{2}Li_m^2$. A slight increase in the interval $PR$ accompanies the peak clipping of $V_2$ and so maintains the area under the curve.

$(C + C_s)$ proceed to a second quarter cycle at the end, $S$, of which they are both completely discharged, having transferred their energy once again to $L$, but with current now flowing in the opposite direction. The standardized energy ensures a standard current.

As $V_2$ passes through zero, $MRZ_1$ becomes forward conducting via $VT_2$ bottomed, and short circuits $L$ and $C$. Because $C$ is discharged at that time, no heavy surge currents flow and no energy is lost. The short-circuit also serves to maintain the current in $L$ at low decrement and therefore holds the standard c.r.t. spot deflexion until the drop, $T$, of $V_1$ switches $VT_2$ off and $VT_1$ on again. $V_2$ now rises to $+E$ as fast as current can be supplied to $LC$ via $VT_1$. During this rise, $T$ to $U$, the change in $i$ is approximately parabolic and in the scan direction but of small amount. From $U$ onwards, $i$ changes linearly with time.

A c.r.t. blanking waveform may readily be obtained from $V_1$.


D. P. FRANKLIN
EMI Electronics Ltd.


### Pulse Stretching in Time-Division-Multiplex

DEAR SIR,—Pulse stretching is now a well-known technique in t.d.m. systems and it is used to enhance the modulating-frequency components at the output of each separated signal channel in the decoder, immediately before the signals are applied to low-pass filters. The principles are dealt with in some text-books, but an adequate theoretical treatment involving both time and frequency domains seems to be lacking. In a recent paper 'A Demonstration T.D.M. System using P.A.M. Techniques' by R. Barrett (March 1965 issue) an attempt is made to derive theoretical expressions. While the principle is sound, this theory lacks detail, and the results are incorrect. The object of this note is to derive expressions for the stretched-pulse p.a.m. signal in both time and frequency domains, using the procedure suggested in Mr. Barrett's article.

Assume that a signal which varies with time according to $f(t)$ is sampled

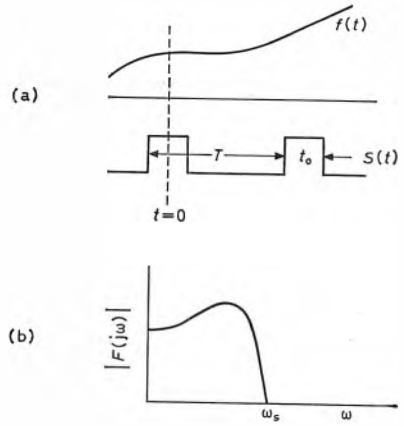



**Fig. 1. Switching function and frequency spectrum**

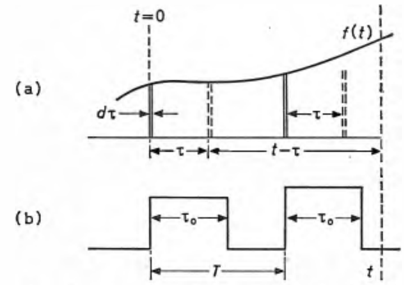



**Fig. 2. Unstretched and stretched pulses**

periodically by a coder at time intervals $T$ sec apart, using the switching function $s(t)$ of Fig. 1(a). The sample pulses have duration $t_0$. The signal $f(t)$ is assumed to have the frequency spectrum $F(j\omega)$ shown in Fig. 1(b), and is limited to $\omega_s$rad/sec. The Fourier spectrum of $s(t)$ is well-known to be

$$\left.\begin{aligned} s(t) &= 1/T \sum_{n\to-\infty}^{n\to\infty} C_n \exp(jn\omega_c t) \\ \text{where}\quad C_n &= kT\,\frac{\sin(nk\pi)}{nk\pi} \end{aligned}\right\} \quad \ldots\ldots\ (1)$$

Here $k = t_0/T$ and $\omega_c = 2\pi/T$

Also the signal spectrum is given by

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

The Fourier spectrum of the p.a.m. signal is then

$$F_m(j\omega) = 1/T \int_{-\infty}^{\infty} \left[ \sum_{n\to-\infty}^{\infty} C_n \exp(jn\omega_c t) \right] f(t)\exp(-j\omega t)dt$$

$$= 1/T \sum_{n\to-\infty}^{\infty} C_n F(j\omega - jn\omega_c) \;\ldots\ldots\; (2)$$

Thus the modulating signal spectrum is shifted from the origin by $n\omega_c$, and weighted by the multiplying factor $C_n$. For $n = 0$, $C_n = t_{o/T}$ and so the modulating frequency spectrum is present, but attenuated by the factor $t_{o/T}$. At $n = 1$ there is a lesser spectrum about the sampling frequency $\omega_c$, and so on.

We now consider the stretching process. First assume that the unstretched sampling pulses are very narrow, as shown in Fig. 2(a). We write $t_o \to d\tau$. Then, in equation (1), $C_n \to d\tau$. The samples are to be stretched for duration $\tau_o$ as shown in Fig. 2(b), and it is seen that the time origin is then located very near the leading edge of a stretched pulse.

Now freeze the observation time $t$ (shown by the dashed vertical line on the right of Fig. 2), and delay the sample pulses by time $\tau$ as shown in Fig. 2(a). The delayed samples, as functions of time $t$, are then given by

$$f_m(t) = d\tau/T \sum_{n\to-\infty}^{\infty} \exp[jn\omega_c(t-\tau)f(t-\tau)] \;\ldots\ldots\ldots\ldots\; (2a)$$

Here we have replaced $t$ by $(t - \tau)$ throughout the right-hand side, but have retained $t$ as the observation time on the left.

For stretching over time $\tau_o$ then, writing the stretched time function of Fig. 2(b) as $[f_m(t)]_s$, and its Fourier spectrum as $[F_m(j\omega)]_s$,

$$[f_m(t)]_s = \int_{\tau=0}^{\tau_o} 1/T \sum_{n\to-\infty}^{\infty} \exp[jn\omega_c(t-\tau)]f(t-\tau)d\tau \;\ldots\ldots\ldots\ldots\; (3)$$

so that

$$[F_m(j\omega)]_s = \int_{-\infty}^{\infty} \int_{\tau=0}^{\tau_o} 1/T \sum_{n\to-\infty}^{\infty} \exp[jn\omega_c(t-\tau)]f(t-\tau) \exp[-j\omega t]d\tau dt \;\ldots\ldots\ldots\; (4)$$

Now write

$$\exp(-j\omega t) = \exp[-j\omega(t-\tau)]\exp[-j\omega\tau]$$

Then all terms in equation (4) involving $t$ have $(t-\tau)$ in common. Further, the order of integration with respect to $t$ and $\tau$ may be reversed since one is held constant during integration with respect to the other. Consider the integration with respect to $t$ first, then. The limits of the integration are $t \to \pm\infty$ so it makes no difference if we replace $dt$ by $d(t-\tau)$, and change the limits to $(t-\tau) \to \pm\infty$. The result is

$$\int_{-\infty}^{+\infty} 1/T \sum_{n\to-\infty}^{\infty} \exp[jn\omega_c(t-\tau)]f(t-\tau) \exp[-j\omega(t-\tau)]d(t-\tau)$$

$$= 1/T \sum_{n\to-\infty}^{\infty} F(j\omega - jn\omega_c)$$

This is evident from equation (2). Also the result is independent of $\tau$ so that equation (4) reduces to the above multiplied by

$$\int_0^{\tau_o} \exp(-j\omega\tau)d\tau$$

Thus

$$[F_m(j\omega)]_s = \tau_o/T \sum_{n\to-\infty}^{\infty} F(j\omega - jn\omega_c) \left[ \frac{\sin(\omega\tau_o/2)}{\omega\tau_o/2} \right] \;\ldots\ldots\; (5)$$

In practical stretching systems then, it is clear that we are only concerned with the modulating-frequency ($n = 0$) part of the spectrum of equation (5). All frequency spectra about the sampling frequency $\omega_c/2\pi$ and its harmonics are attenuated and distorted by the term

$$\frac{\sin(\omega\tau_o/2)}{\omega\tau_o/2}$$

There are two limiting conditions of practical importance. One is the condition $\tau_o = T$ in which case the stretching process becomes the zero order hold of sampled data systems. The other condition is where sampling is at the minimum rate which will satisfy the Nyquist theorem, i.e. $\omega_c = 2\omega_s$. The modulating spectrum is then

$$F(j\omega)\ \frac{\sin(\pi(\omega/\omega_c))}{\pi(\omega/\omega_c)}$$

and at $\omega = \omega_s$ this becomes $(2/\pi)F(j\omega)$.

Hence, with sampling at very near the Nyquist rate, pulse stretching to the zero-order-hold extreme has the effects of

(1) Increasing the separated modulating-frequency decoder signal amplitudes from $t_o/T$ to near unity at very low modulating frequencies.

(2) Introducing frequency distortion in the modulating frequency output according to the curve

$$\frac{\sin(\omega T/2)}{\omega T/2}$$

and thereby requiring an equalizer to be inserted, after stretching, which has a zero boost at d.c. and boost of $\pi/2$ at $\omega = \omega_s$.

(3) Attenuating and distorting the frequency spectra about the sampling frequency and all harmonics of the sampling frequency.

Equations (3) and (5) are the time and frequency-domain expressions for the stretched samples.

Yours faithfully,

R. Kitai,
McMaster University,
Hamilton, Ontario.

**The author replies:**

Dear Sir,—I agree with Dr. Kitai that there is an error in equation (6) of my article (March 1965 issue), but the error does not affect the results obtained after passing the stretched samples through the demodulating low-pass filter, which are in agreement with those derived by him.

Equation (5) of Dr. Kitai's letter should include a multiplying term $\exp\left[\frac{-j\omega\tau_o}{2}\right]$ resulting from the last integration and representing the delaying effect of pulse stretching. I should also point out that the second sentence after equation (2) should start . . . "For $n = 0$, $C_n = t_o$ and so . . .", and equation (2a) should read

$$f_m(t) = d\tau/T \sum_{n\to-\infty}^{\infty} \exp[jn\omega_c(t-\tau)] \,.\, f(t-\tau)$$

Yours faithfully,

R. Barrett,
Lanchester College of Technology,
Coventry.

## Photomultipliers Fatigue

Dear Sir,—We note with interest the remarks on fatigue in photomultipliers by R. J. Thresher and E. Gordy[1,2]. For many years, we have used EMI photomultipliers with anode currents in the range $10^{-12}$ to $10^{-6}$A for absolute measurements of radiation from upper atmosphere seeding experiments[3], and have found fatigue to be a problem at these levels. Since our observations are associated with rocket trials, the equipment often stands idle for many weeks at a time. After such a period of unpowered dark storage we condition the photomultipliers by running them (still in darkness) for about twelve hours, to stabilize the dark current. The tubes are then exposed to a light which gives an anode current in the range $10^{-10}$ to $10^{-6}$A for 30 to 60min. During this time the anode current drops by about 50 per cent despite constant illumination. This produces a condition in which stable measurements can be made for many tens of minutes. Our feeling is that the tubes are then in a mildly fatigued condition.

Yours faithfully,

J. R. V. Groves, C. H. Low,
Weapons Research Establishment,
Australia.

REFERENCES

1. *Electronic Engng.* 37, 120 (1965).
2. *Electronic Engng.* 37, 339 (1965).
3. *J. Sci. Inst.* 42, 430 (1965).

**The correspondent replies:**

Dear Sir,—Messrs. Groves and Low have produced further support for the existence of photomultiplier fatigue at $10^{-6}$A and it is most interesting to note that the effect is experienced down to $10^{-10}$A.

One possible way of counteracting this effect would be to arrange for the photomultiplier to have frequent looks at a light standard and to arrange for the tube gain to be stabilized during the look period by resetting the tube h.t. This could be done automatically.

Yours faithfully,

R. J. Thresher,
Harrow Weald.

# NEW BOOKS

### Electromagnetic Diffraction and Propagation Problems

By V. A. Fock. 414 pp. Med. 8vo. Pergamon Press. 1965. Price £7

THIS book is Volume 1 of the International Series of Monographs on Electromagnetic Waves published by the Pergamon Press and of which the author is a co-editor. It is actually a collection in their original form of papers by the author and his colleagues in the Soviet Union published during the past twenty years and, as he admits in the Introduction, this has led to some unavoidable repetitions. It is interesting to see how the work developed from stage to stage as the problem was gradually generalized, but from the general reader's point of view it would have been more helpful if the author had taken the opportunity to use his material to write a text-book that would have been much shorter and more assimilable. However, we must be glad to have in English this important and original work much of which was originally only available in Russian.

The author is well known for his contributions to the basic theory of the diffraction of waves round convex surfaces with its special reference to the penumbral transition from the lit to the shadow region in the classical problem of propagation over a spherical earth. It is thus natural that this aspect of the subject has a prominent place. Part I of the book, which is called 'asymptotic theory of diffraction' takes the paraboloid of revolution as a mathematical model leading on to generalized forms of the Fresnel laws of reflection for a surface of arbitrary form. Part II on 'tropospheric radiowave propagation' is concerned mainly with the effect of an inhomogenous atmosphere and the conditions for superrefraction.

This is not a book for the practical engineer as the approach is wholly mathematical at an extremely high level. Indeed only those who are already well acquainted with the theory of ground-wave propagation are likely to make any headway with it. Some of the longest and most difficult chapters have no diagrams or calculations of any sort, and often the physical significance of what is being argued out only appears at the end with the results left in a complex mathematical form. Nevertheless the work is full of analytical techniques that should be a stimulus and a fund of ideas for research workers in wave theory.

Perhaps the most useful part of the book is the treatment of superrefraction in the last three chapters where special attention is paid to the horizon region where the electromagnetic energy gets into a duct with the conditions under which the waveguide becomes leaky at sufficiently low frequencies. The treatment is based on a model of refractive index as a function of height that leads to a solution in terms of parabolic cylinder functions and some numerical and graphical results are given. They are mainly illustrative of the type of answer that can be obtained with sufficient computational effort rather than the provision of curves that can be applied by engineers to a specific problem.

The author was one of the first to make extensive use of Airy functions in diffraction theory and in an appendix a detailed discussion of the functions and their derivatives is given with tables for real values of the argument between $-9{\cdot}00$ and $9{\cdot}00$ with an interval of $0{\cdot}02$ suitable for linear interpolation for most purposes. The translation is excellent, and allowing for a rather long errata published as an insert, the typesetting of the complicated mathematics is commendable. The high price is an acknowledgement that the book will have a very limited sale, but it should be in all appropriate technical reference libraries and will be well received in the United States of America where so much has been done to make the work of the author known outside the Soviet Union.

G. MILLINGTON.

### Fundamentals of Semiconductor Devices

By J. Lindmayer and C. Y. Wrigley. 486 pp. Med. 8vo. D. van Nostrand Co. Inc. 1965. Price 93s.

THIS book certainly lives up to the authors' expressed intention of providing an unconventional approach to the explanation of semiconductor devices. This unconventionality has confused the reviewer slightly in deciding at whom the book is aimed. It seems best suited to circuit engineers who have been more familiar with valves and are now approaching transistors for the first time (are any of these left still?). On the other hand, some words in the preface and the provision of examples indicate that students in universities or technical colleges may be the target. In the latter case, it seems likely that only a few will find that this is just the textbook that makes everything about semiconductor devices crystal clear, while all other books are impenetrable.

Having said this, it must be readily admitted that there is certainly no case against (and perhaps one for) introducing some of the circuit properties of transistors, before diving into the details of semiconductor physics. Since this latter is the field of the reviewer, unusual statements in this area have been more obvious than in other areas. For instance, the authors express very briefly a somewhat restricted view of phonons, and omit any mention of the existence or properties of compensated semiconductors. Moreover, it is implied that the drift field in the base of mesa and planar transistors is always of significant importance.

As far as the organization of the book is concerned, the major criticism is an almost complete lack of names and references, which considerably reduces its value as a textbook. The index is limited, and cross-references in the text are not as good as they might be. There appears to be at least one small and not very important misprint, which can no doubt be left as an additional exercise for the reader.

P. C. NEWMAN.

### Field Effect Transistors

By L. J. Sevin. 138 pp. Crown 4to. McGraw-Hill Book Co. 1965. Price £4.

THIS book is well prepared and laid out. The quality of the printing and the diagrams is superb but this is reflected in the rather high price for a book of this nature and size (130 pages).

Although the book mentions m.o.s. transistors it deals exclusively with the applications of unipolar or junction field effect transistors. It is admitted at the end of the book that it was written too early to really assess the insulated gate transistor. Perhaps this should be made clearer in the title pages so as not to mislead people whose interest in f.e.t.'s is stimulated by the insulated gate device.

In dealing with the junction transistor the book is straightforward and easy to read. The mathematics is all that is necessary, is well presented and does not contain unexplained jumps to confuse the reader.

The circuit examples given are described simply, both mathematically and experimentally. A minor criticism is that both the unipolar and bipolar active devices are given in most cases as type numbers and will tend to date the book as new and better types become available, especially as the book does not make it clear what parameters were used to select the given transistors. The similarities between the valve and the junction f.e.t. are made very clear and thus the book shows how existing valve theory and circuits can be adapted for semiconductor use.

Although both the n and p channel devices are described the book does not deal with complementary circuits. Some confusion exists about the symbol used, particularly in the direction of the arrow on the gate lead and no clear convention seems to have been adopted. In the initial circuits shown an outward facing gate arrow is associated with a negative drain voltage. In some later circuits the same symbol is used with positive drain supplies. Towards the end of the book the complementary symbol with an inward facing arrow is used also associated with a positive drain supply voltage, together with a symbol having no arrow at all.

With the reservation that the book does not deal with the m.o.s. and not specifically with the planar junction f.e.t. it can be recommended as a text book on the theory and application of unipolar field effect transistors.

R. E. HAYES

## Field and Wave Electrodynamics

**By C. C. Johnson. 456 pp. Med. 8vo. McGraw-Hill Book Co. 1965. Price £5**

ALTHOUGH a number of texts exist which treat part of the subject area of this book, the author has made a significant contribution with the present work, the objective being to present a unified approach to electrodynamics. A review of electromagnetic laws is followed by a resumé of those mathematical techniques which are required in the solution of boundary value problems. Naturally, only a limited treatment is provided here, but the presentation is easy to follow and provides a useful introduction to subsequent chapters. The next group of chapters consider unbounded field systems, transmission systems, waveguide discontinuities and cavities. Although these topics have been extensively considered before, the author provides us with a fresh view. There are many clear diagrams and graphs and a good balance has been achieved between elementary and advanced concepts. Of particular mention is the clear exposition of phase, group and signal velocities, also the section on waveguides can be of particular help to undergraduate students. Chapters on periodic structures, electromechanical waves on electron beams and coupled transmission systems cover material not generally to be found in a graduate text on electromagnetic theory. The author has made research contributions in this field and the material presented will be especially welcomed by those commencing work on microwave electron beam devices.

The inclusion of a chapter on optics reflects the increasing interest of microwave engineers in lasers and optical systems. An electromagnetic approach is adopted and a review of basic phenomena is presented which includes propagation through isotropic and anisotropic crystals. Brief mention is made of propagation of light through the atmosphere.

A comprehensive review of electromagnetic interaction with gaseous plasma is given which includes waves in bounded and unbounded regions. Brief mention is made of electron beam interaction with plasma but the subject is too big to be considered in two pages. The concluding chapter on ferrites contains some useful ideas but the treatment does not do justice to this important subject area. For example, the author presents an impression that the Faraday rotation effect is of considerable use in devices, a viewpoint which is almost ten years out of date.

Each chapter contains a reasonable number of problems, some are quite searching, others are rather trivial while, generally, it is to be regretted that more do not contain implied answers. A number of good references are provided with each chapter, these, perhaps rather naturally, apply mainly to the United States literature.

In summary, the author has treated a number of conventional topics in an enterprising way and has considered most attractively more specialized topics of interest to those concerned with electron beam and plasma devices. The book can be well recommended as advanced reading at final year undergraduate level in the United Kingdom and as a reference work for those starting research.

P. J. B. CLARRICOATS.

### British Miniature Electronic Components Data 1965-66

**By G. W. A. Dummer and J. Mackenzie Robertson. 984 pp. Pergamon Press. 1965. Price £9**

In the fourth edition of this book 90 per cent of the components described are appearing for the first time. Details of components previously covered and still in production are contained in the third (1963-64) edition.

A feature of the new edition is the inclusion of dimensional drawings.

### Nucleonic Instrumentation

**By C. C. H. Washtell and S. G. Hewitt. 144 pp. Demy 8vo. George Newnes Ltd. 1965. Price 42s.**

This book has been written to cover the function, performance and circuit techniques of the wide range of instruments employing electronic circuits. A non-mathematical treatment is followed throughout, with the intention of giving to the instrument user rather than the designer a basic understanding of circuit functions, instrument uses and maintenance.

### Guida Industriale Commerciale Di Elettronica 1965/1966

**By P. G. Portino. 748 pp. Demy 4to. Edizioni Minerva Tecnica. 1965. Price L150,000**

This buyers' guide, which is printed in Italian, English, French and German, contains a list of Italian manufacturers of electronic components and equipment, together with agents in Italy acting for foreign firms.

There is also a four-language dictionary of technical terms.

### Progress in Semiconductors—9

**Edited by A. F. Gibson and R. E. Burgess. Royal 8vo. Temple Press Books Ltd. 1965. Price 80s.**

Volume 9 in this series contains five papers dealing with various aspects of semiconductor research.

### Localisation des Satellites Basée sur le Principe de L'Interferometre Radioelectrique—Volume One, Part 1 and Satellites et Fusées Porteuses Lances Depuis Spoutnik 1, Part 2

**By G. Nachszunow. 408 pp Part 1. 136 pp Part 2. Med. 8vo. Madame Willy Nachszunow. 1965. Price 45F.**

The object of this book is twofold, to describe the theory and principles for the construction of a radio-interferometer and the practical applications of this equipment to the Minitrack Mark I and II.

Part 1 of this book contains eight chapters starting with the early history of satellites dating back as far as 1840 when the idea of a satellite was suggested by Permit in France.

A ninth chapter is added which consists of a separate book (Part 2) and contains full technical information on all the satellites launched into outer space from Sputnik 1 on 4 October 1957 to Titan 3C on 18 June 1965.

### Conference Proceedings, Vol. 26, Part II Fall Joint Computer Conference

**By A.F.I.P.S. 144 pp. Demy 4to. Macmillan. 1965. Price 37s. 6d.**

Volume 26 contains the papers presented at the American Federation of Information Processing Societies' Conference dealing with Very High Speed Computer Systems.

### Detection Theory

**By I. Selin. 128 pp. Demy 8vo. Oxford University Press. 1965. Price 40s.**

This book deals with the theory covering the testing for the presence of a desired process (signal) in the presence of an undesired random process (noise) and to applications in radar, communications and control.

Included in the book are such subjects as the complex representation of real signals, integral equations with symmetric kernals and the Karhunen-Loeve expansion.

### Electrical Insulation Measurements

**By W. P. Baker. 180 pp. Med. 8vo. George Newnes Ltd. 1965. Price 50s.**

This book is intended to assist the industrial laboratory worker in the measurement of electrical insulation and to guide the supply engineer in making the best deductions from his more limited tests.

The essentials of the physical and chemical background are presented so that simple correlations between electrical performance and molecular structure can be observed.

### The Story of the Laser

**By J. M. Carroll. 181 pp. Demy 8vo. Souvenir Press Ltd. 1965. Price 21s.**

This book describes in non-technical language what the laser is and explains the more important applications in the military, space exploration, data processing and medical fields.

### Basic Electronics A Programed Course in Circuits

**By RCA Institutes Inc. 534 pp. Med. 8vo. Prentice-Hall International. 1965. Price 68s. 6d.**

This book is an introduction to electronic circuits for the beginner with no previous knowledge or experience in the subject.

### Recent Advances in Selenium Physics

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

This book contains the papers presented at the Symposium on Solid and Liquid State Selenium Physics arranged by the European Selenium-Tellurium Committee and held at the Chemical Society, London, in June 1964.

# NEW EQUIPMENT

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

(Voir page 57 pour la traduction en français; Deutsche Übersetzung Seite 64)

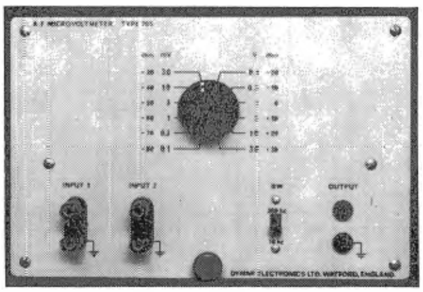
### A.F. MICROVOLTMETER

**Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire**

(*Illustrated below*)

The latest addition to the Dymar System of plug-in instruments is the type 705 a.f. microvoltmeter. This is a sensitive valve-voltmeter designed to measure a.c. voltages in the audio-frequency range, although its frequency response extends to 300kc/s with full accuracy.

A feature of this instrument is that it has a 'balanced input' which permits direct measurement of push-pull voltages with respect to earth.



The 'in-phase' rejection over most of the frequency range is greater than 60dB allowing the measurement of balanced voltages in the presence of much larger interfering voltages. Either input can be used for measurement of unbalanced voltages with respect to earth.

Input sensitivity is 100$\mu$V for full scale deflexion with an inherent noise level of less than 20$\mu$V: this noise level can be reduced to 8$\mu$V by restricting the bandwidth to 10kc/s.

The input impedance on each input is 10M$\Omega$ and output terminals are provided permitting the instrument to be used as a sensitive pre-amplifier for driving oscilloscopes or power amplifiers.

**EE 89 751** *for further details*

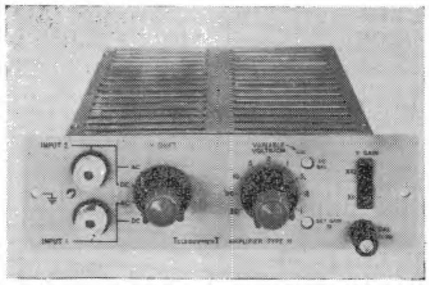
### WIDEBAND AMPLIFIER

**Telequipment Ltd, 313 Chase Road, Southgate, London, N.14**

(*Illustrated above right*)

A new unit in the range of plug-in amplifiers by Telequipment Ltd is the type H, designed for wide bandwidth requirements and particularly suitable for rise time measurement.

There are two input sockets with 60dB isolation between inputs, which are switch selected. Bandwidths are d.c. to 25Mc/s (−3dB) with sensitivity of 100mV/cm to 50V/cm, and d.c. to 5Mc/s



(−3dB), sensitivity 10mV/cm to 5V/cm. Sensitivity in each case is switched in 1 : 2 : 5 ratio, and a gain control gives continuous variation between fixed step attenuator ranges.

The type H amplifier, in common with the other five plug-in amplifiers produced by the company, is fully interchangeable between Telequipment oscilloscopes types S43, D43 and D43R, as well as the new D53, which has just been introduced by the company.

**EE 89 752** *for further details*

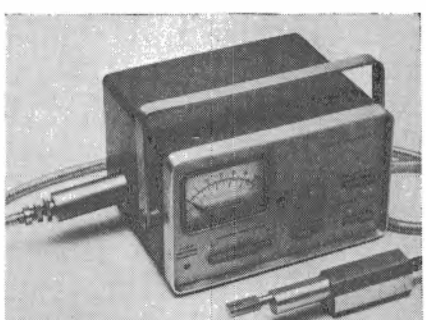
### SURFACE FINISH TESTER

**Payne Products International Ltd, Buckingham Avenue, Trading Estate, Slough, Buckinghamshire**

(*Illustrated below*)

Payne Products International Ltd, producers of Lapmaster flat lapping machines, announce a new microtester—the Mark II Diavite—capable of measuring all recognized ranges of surface finishes. Working on 110V to 220V a.c. single phase, the new model is capable of measuring centre line averages between 0 and 300$\mu$in. This scale is divided by means of a push-button into four sections, up to 10, 30, 100 and 300$\mu$in.

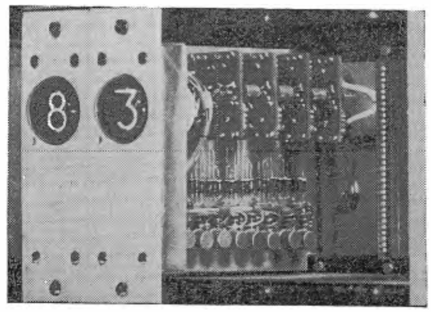
The instrument has two cut-off distances, ·003in for very fine concave or convex profile sections, or ·03in for flat profile sections.



New features include a tolerance light which may be set to the limit values for c.l.a. or $R_{max}$ allowed on any component to be tested. If the limit is exceeded a red warning light on the face lights up.

Another new feature is the incorporation of three stylus strokes of varying length, 0·4, 0·8 and 0·16in. The stylus speed is ·05in/sec. The new model is portable weighing only 13·31b.

**EE 89 753** *for further details*



### DECADE COUNTER MODULES

**Quarndon Electronics Ltd, Slack Lane, Derby**

(*Illustrated above*)

This range of decade counter modules is intended for industrial systems and laboratory use where the expense of a complete electronic counter is not justified. These modular units offer a highly flexible system for specialist applications.

The units occupy a front panel area of 4¾in by 1¼in, and are 5¼in deep, with terminations filling the standard 32-way printed circuit board connector. The count is displayed on a standard type numerical display tube.

Units are available made with either silicon or germanium semiconductor devices. The maximum counting speed for the DCM 501 unit is 1Mc/s. Other germanium units are available with counting speeds of 5Mc/s and 12Mc/s. The silicon units have counting speeds of 200kc/s, 2Mc/s and 30Mc/s. Reversible counters are also available. The decimal point is provided by a miniature neon bulb which is operated externally.

The input pulse required depends to some extent on its rise time. For maximum speed a 2½V pulse with a rise time of 500nsec is required.

The output of the unit is sufficient to drive a further unit so that a multidigital counter can be built up. 8-4-2-1 binary coded decimal outputs are available for remote display, etc.

The unit may be reset at zero by the application of a reset pulse to the reset terminal. Crystal frequency standards, gating units and remote display units are available to complete the range so that complete counters can be built.

**EE 89 754** *for further details*

## RUGGEDIZED REFLEX KLYSTRONS

**The English Electric Valve Co. Ltd, Chelmsford, Essex**

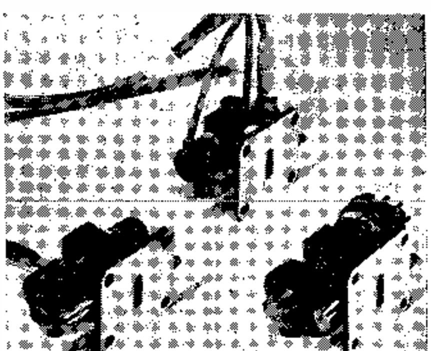
*(Illustrated below)*

The English Electric Valve Co. Ltd has produced a new series of compact X Band reflex klystrons to meet the expanding demand for relatively inexpensive, fully ruggedized local oscillators, having a wide range of frequency/temperature coefficients.

This latest group of tubes embody the same stringent production techniques as those employed in the construction of ruggedized EEV klystrons intended for military use. The obvious advantages that stem from 'building-in' such experience are immediately apparent in the enhanced performance and reliability achieved with the new tubes.

A feature of this series of klystrons is the ability to choose one with a frequency/temperature coefficient suited to the particular application. The K3007, for example, has a coefficient of −130kc/s/°C, which is designed to match that exhibited by most commercially available X-band magnetrons. Coupled with this, a broad degree of manufacturing interchangeability exists between the new tubes, permitting the ready permutation of performance parameters to suit individual requirements for non-standard klystrons.

Typical of the tubes included within this series is the K3003, which is electrically identical to its flying lead variant the K3020. Ideally suited for marine use, the excellent frequency drift characteristics of the K3003 are of particular note, having remained within a 10Mc/s variation during testing of over 3 000 hours. In addition to this, the peak frequency modulation of the tube is held to a specification maximum of 200kc/s under vibrational loadings of 10*g*.



**EE 89 755** *for further details*

## MICROLOGIC TUTOR

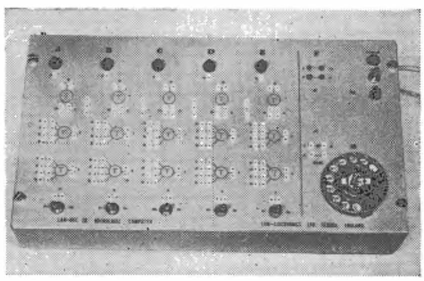
**Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire**

*(Illustrated below)*

The LD-20 is a development of the well established LD-15 and employs semiconductor integrated circuits (s.i.c.) for the logic. Using these modules electronic engineering becomes a problem of interconnexion between circuits rather than the detailed design of the individual circuits. The Micrologic Tutor gives the potential user of s.i.c. the opportunity to gain practical experience in all aspects of work with integrated digital circuits. Each s.i.c. consists of a complete digital, or linear, circuit in which all the components are fabricated in a single small semiconductor slice, packaged in a transistor can (such as the TO-5 size) or in a flat pack (typically measuring 6·6mm square by 1·3mm thick).

The s.i.c. units in the LD-20 operate from a supply line of only 3·6V, at a current of less than 1mA, and this is typical of the trend toward low level logic. Obviously, these low power levels are of great advantage in designing compact equipment.

Fifteen s.i.c. gates are available in the



LD-20 with five additional two-stage buffer amplifiers to drive the indicator lamps. The LD-20 also gives greater versatility than the LD-15 since five double-throw toggle switches are provided for input purposes, in addition to the telephone dial. These switches give optional positive or negative voltages at adjacent sockets for patching into the logic. Although the logic circuits are designed bascially as NOR gates (with 0V = 0 and positive supply voltage = 1), by inverting the sign conventions these fifteen gates may be used for the NAND function simply by operating the input toggle switches.

The Micrologic Tutor incorporates a mains-operated power supply to give the correct d.c. voltages and will be supplied with the comprehensive LAN-DEC Digital Educational Computer Handbook (over 100 pages), the Micrologic Handbook (prepared by S.G.S.-Fairchild) and a short Handbook written specifically for the LD-20 Micrologic Tutor.

**EE 89 756** *for further details*

## STABILIZED E.H.T. UNIT

**Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex**

*(Illustrated above right)*

One of the first units to be produced by the newly-formed Precision Measure-

ments Division of Cathodeon Electronic Ltd, is the PMD 101 5kV stabilized power supply.

The unit gives either positive or negative outputs in relation to earth at any pre-set current up to the full rating of 5mA. The pre-set control is on the front panel.

Full overload protection is by a specially designed transistorized module, adjustable to respond over the current range of 2 to 5mA. Short circuit current is only 20 per cent greater than the overload threshold setting.

Regulation of the supply unit is better than 0·1 per cent over the full voltage range and the twin-scale meter can be switched to read either current or voltage.

The PMD 101 has applications in such fields at photomultiplier supplies and e.h.t. supplies for cathode-ray tubes.

The PMD 101 is supplied in a cabinet for bench use and is also available for standard 19in rack mounting.

**EE 89 757** *for further details*

## TEMPERATURE CONTROLLER

**C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5**

*(Illustrated on page 50)*

C.N.S. Instruments Ltd is producing a new temperature controller. This instrument designated the S.R.2 is a fully transistorized, high precision, proportional temperature controller employing saturable reactors.

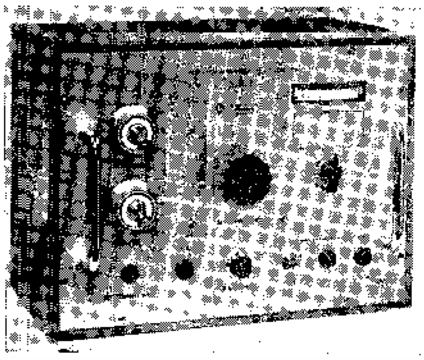
A number of novel features are incorporated; these include current limiting facilities (corresponding to a power ratio of 30 : 1) thus permitting platinum and molybdenum furnaces to be controlled.

A changeover switch allows the metering of either the proportion of power output from the controller or the out-of-balance voltage of the platinum resistance thermometer bridge to be read, and thus the temperature deviation from the set point may be estimated therefrom. Controls are fitted to allow repetitive tests at one temperature with extremely small overshoot.

Printed circuit board construction allows rapid replacement of the majority of the entire circuit in the unlikely event of component failure.

It is suitable for use with resistive or inductive loads and the rugged saturable reactor, which controls the actual furnace power, is able to withstand excessive voltage transients and current overloads.

The S.R.2 incorporates the C.N.S. 50-turn precision slide wire potentio-



meter which may be reset to one part in 5 000, thus allowing extremely accurate temperature setting.

Tests carried out on typical smaller furnace installations have shown that under normal conditions from an unstabilized mains supply the stability of control is better than $\pm 0{\cdot}02$ per cent at 1 000°C when using a C.N.S. platinum resistance thermometer. Automatic mains voltage compensation is incorporated responding within a few cycles to mains supply variations.

The small power dissipation (40VA) of the controllers allows close stacking without excessive overheating occurring and hence no forced cooling is required.

**EE 89 758** *for further details*

### DIGITAL VOLTMETER

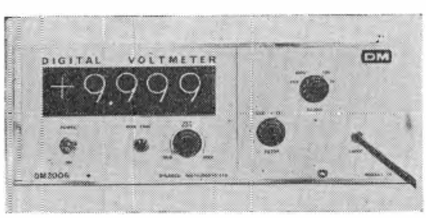
**Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire**

*(Illustrated below)*

The Dynamco Instrument DM2006 is an all solid state digital voltmeter using a new voltage conversion technique giving unusual economy of circuit components. This enables Dynamco (previously Digital Measurements Ltd) to produce a modestly priced voltmeter without sacrificing high performance and construction standards.

The instrument consists of a basic measuring and display unit (scale of 9999) with a plug-in module to determine the ranging facility required. The simplest plug-in unit has four ranges from 100μV to 1kV and gives a reading accuracy of 0·05 per cent. Read-out signals to drive output devices can be provided using plug-in interchangeable cards, without detriment to the high common mode rejection of the instrument. It is planned that modules giving higher sensitivity, automatic ranging, a.c. measurement, and resistance measurement will become available as the development programme progresses.

Considerable emphasis has been placed on reliability and, for example, the DM2006 is constructed using epoxy-glass printed circuit cards, silicon transistors and military type approved components throughout.




**EE 89 759** *for further details*

### TWIN LOW-PASS FILTER

**The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey**

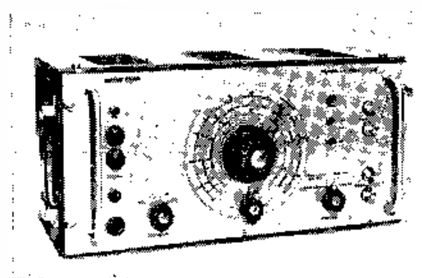
*(Illustrated below)*

A new l.f. filter by Wayne Kerr consists of two identical channels which can be used independently, in cascade, or in a band-pass arrangement. A single tuning control operates over 28 ranges to permit both sections to be set precisely to any desired frequency between 0·1592c/s and 1592c/s.

As twin low-pass filters, the SA500 is particularly suitable for measurements on the performance of control systems. One filter is inserted in the output lead of the system to improve the signal-to-noise ratio. The second filter is inserted in the reference signal path to introduce the identical characteristic and so leave the overall transfer function of the system unaffected by the filter.

With the two low-pass filters in cascade the attenuation rate is 36dB/octave. In the band-pass condition there is unity gain and zero phase-shift at the centre frequency and an attenuation of 18dB/octave either side.

The input impedance of each channel is 2MΩ (resistive) and the output impedance is less than 60Ω. The SA500 is available for 19in rack mounting or in a portable case.



**EE 89 760** *for further details*

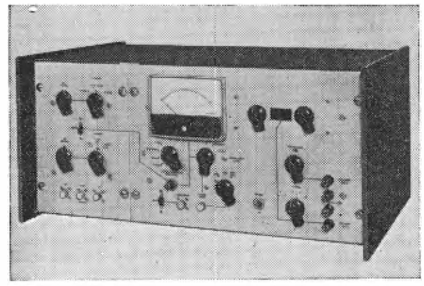
### DISTORTION MEASURING EQUIPMENT

**Standard Telephones & Cables Ltd, S.T.C. House, 190 Strand, London, W.C.2**

*(Illustrated above right)*

Introduced to make dynamic sweep measurements on microwave and other telecommunications systems, the 74834-A distortion measuring equipment measures the amplitude and phase distortion of the transmitted signal caused by non-linearity in the system. In conjunction with a suitable oscilloscope it will measure differential amplitudes of 0·05 per cent to 100 per cent and differential phase from 0·02° to 90°.

Four alternative h.f. test signals are available: one of these is at the television colour sub-carrier frequency of 4·4297Mc/s and the other three at 304 kc/s, 2Mc/s and 7Mc/s. Other frequencies can be provided to satisfy particular requirements. Any one of the h.f. test signals can be added to the l.f. sweep signal, and the amplitude of the latter



can be pre-set so that the composite signal will sweep in turn over various working amplitude ranges of the system. The sweep frequency is 81c/s. Markers are incorporated in the oscilloscope display to facilitate reading of the amplitude and phase distortion.

The equipment can be used alone for loop measurements, or with another similar equipment for end-to-end measurements; alternatively, the receiver can be used at one end, and the generator can be removed from the case and used at the other end in conjunction with an extra power and monitoring unit which is available as an accessory.

**EE 89 761** *for further details*

### REFLEX PHOTO-ELECTRIC UNIT

**Lancashire Dynamo Electronic Products, Rugeley, Staffordshire**

*(Illustrated below)*

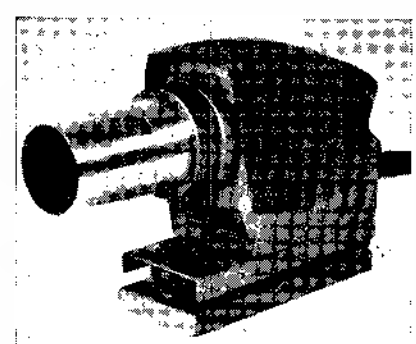
LDEP (MI Group) has extended its standard range of photo-electric equipment with the introduction of a new head unit series RPH 6, for use with its transistorized control equipment.

The RPH 6, which operates on the reflex principle, consists of a single remote photo-electric head unit housing both the 'projector' and 'receiver' components. These components are so arranged that the receiver cell will 'see' the projected light beam reflected from the target object.

The equipment is suitable for all applications involving positioning, counting, over-travel protection, and other similar requirements.

The head unit is of light alloy pressure die cast construction and comprises two main components, the body and cover. The detachable cover allows easy access to the interior without disurbance of the optical components. A rubber gasket ensures a weatherproof seal and the overall design enables rigid mounting to be obtained, eliminating misalignment from shock and vibration.

The reflex assembly consists of a stan-



dard 8W light source to which has been added a 1¼in diameter × 1¼in focus glass reflector and cell holder.

The reflector is mounted in an aluminium mount and provision is made for adjustment in vertical, horizontal and front to rear planes to ensure sharp definition of filament image from approximately 20in to the maximum operating limit.

For short range work or for conditions where a smaller and more sharply defined filament image is required, an additional 1¼in diameter lens is fitted.

By using various optical and electrical components, which can be assembled in a variety of combinations to suit particular site conditions, the RPH 6 will operate over distances from 1in to 20ft approximately.

The basic unit is fitted with an ORP.60 semiconductor photocell which varies its resistance on exposure to light. Operation is satisfactory at ambient temperatures up to 45°C.

**EE 89 762** *for further details*

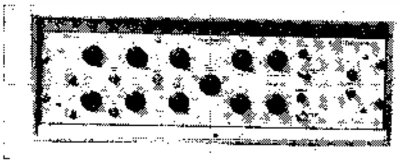
### PULSE GENERATOR

**K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4**

*(Illustrated below)*

The pulse generator model T-15, is an all solid state instrument which has been designed to be extremely compact utilizing printed circuit boards and high quality components. Although compact this instrument is very versatile; it provides single or double pulses with simultaneous positive and negative polarity. Pulse widths and delays from 30nsec to 30msec are available and the internal oscillator has a frequency range from 15c/s to 15Mc/s. Pulse widths up to nearly 100 per cent duty cycle can be obtained. Positive and negative pulses having an amplitude up to 10V peak into 50Ω are obtainable with continuous control of amplitude.

The unit can be triggered from its own internal oscillator, internal single shot push-button, or from an external waveform. Gated and sweep frequency operation can also be used from external signals.



**EE 89 763** *for further details*

### RELAY DELAY UNITS

**Distributed by: D. Robinson & Co. Ltd, Victoria House, 44 Park Street, Camberley, Surrey**

*(Illustrated above right)*

Clifford & Snell Ltd has announced two newly developed electronic delay units for use with the D2600 plug-in relay system.

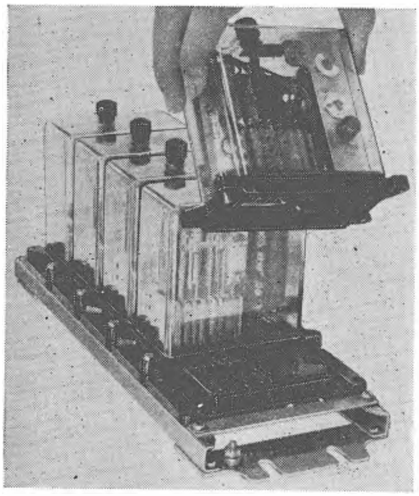
The first of these is a version of the 'Superslug-A' for operation from 110V a.c. As can be seen from the illustration, it is externally identical to a D2600 relay and plugs into the same type of socket.

It is called a Superslug because it 'slugs' or delays the release of an associate relay in a similar manner to a built-in copper ring, but does so for a much longer period.

It employs a cold cathode valve fed from a voltage doubler circuit and can give delays from 1 to 72sec with a repetitive accuracy of the order of ±1 per cent.

It is made for reliability, and is employed in the type of application where Clifford & Snell relays are used, e.g. ships, power stations etc. Precautions include an internal fuse and a limiting resistance.

The second new unit is the 'Superslug-B2' which will operate from both 24V d.c. and the 50V d.c. by altering a connexion. It supercedes the 'Superslug-B' being of improved design and is about a third less in price.



**EE 89 764** *for further details*

### DATA TEST SET

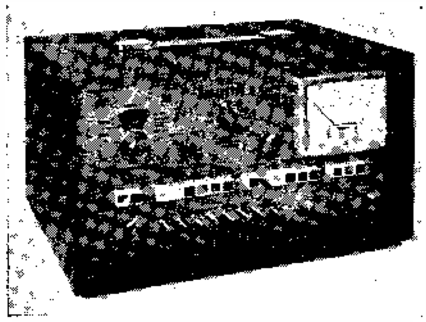
**The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex**

*(Illustrated above right)*

This new instrument, type L.675, has been developed by the Telecommunications Division of M.E.L. to meet the demand for a portable device to measure the performance of data transmission equipment and circuits. It has been designed to meet the G.P.O. specification for the Datel Tester 1A.

The instrument incorporates separate transmitting and receiving sections to allow transmission to be effected down one channel of a two-way circuit while receiving on the other channel at a different speed.

Features of the data test set are a large unambiguous direct reading distortion meter for measurement of mean bias distortion and peak value of random distortion, an error counting facility and a speed range which permits operation down to normal telegraph speeds. The receiving speed range maximum is 5 000 bits/sec which covers any foreseeable data transmission speeds.



Exceptional accuracy makes the instrument suitable as a laboratory tool for manufacturers of data transmission equipment and computer manufacturers whose machines embody input/output facilities for data transmission.

The instrument is automatically protected against inadvertant connexion to a mains supply of wrong voltage. On disconnexion, the mains voltage adjustment automatically reverts to its highest voltage tapping.

The L.675 and its stabilized power supply is housed in a rugged metal case measuring only 16 × 11 × 8¾in. It is fitted with a carrying strap, and the front panel, control knobs, meters and indicators are protected by a hinged front cover which stows in the case when the instrument is in use.

**EE 89 765** *for further details*

### PRECISION D.C. MOTORS

**Distributed by: Leland Leroux Ltd, 145 Grosvenor Road, Westminster, London, S.W.1**

*(Illustrated below)*

These miniature precision d.c. motors, manufactured by Brion Leroux & Cie., of Paris and designed essentially for precision measurement and integration, for servo mechanisms, remote control and as tachometric generators, are now available in the United Kingdom through Leland Leroux Ltd.

These motors are made in two basic mechanical sizes, and in each case are available in several different types. They have a speed in precise proportion to the voltage applied to their windings; or when used as generators, produce a voltage proportional to the rate of revolution. They have a very high efficiency, low starting current, very low inertia and time-constant. A range of reduction gearboxes is available for each model.

The smaller model—the Microtor—is available in two types: Type 1 gives

3 000 rev/min with 5V, and requires a starting power of 100$\mu$W; Type II gives 3 000 rev/min with 10V and requires a starting power of 160$\mu$W. Gearboxes with reduction ratios of 1:18; 1:50, and 1:150 are available for this model.

The larger model—the Birotax—has two independent windings which may be used in series or separately (one as a motor, one as a generator). To give 3 000 rev/min Type I requires 50V on each winding; Type VI 20V on each winding; Type V, which is of specially low resistance designed for transistor circuits, requires 5V on each winding; and Type II is of a special type with two different windings—one for 100V as a motor and the other 9·5V to serve as a tachometric generator. Gearboxes with reduction ratios of 1:18, 1:50, 1:150, 1:2 000 and 1:3 000 are available for this model.

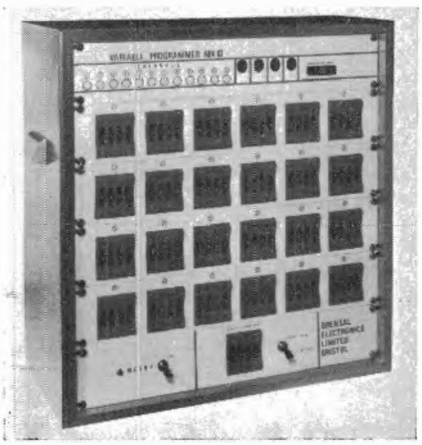
**EE 89 766** *for further details*

### VARIABLE PROGRAMMER

**Brensal Electronics Ltd, Charles Street, Bristol, 7**

*(Illustrated below)*

Called the Variable Control Programmer Mk III, this instrument is completely solid state throughout and



represents a new concept in design of this type of control instrument.

The programmer is intended for machine tool numerical control or process control applications where the need for carrying out rapid changes in a complex switching programme frequently arise. One of the advantages of this instrument is that any change may be effected in the programme without in any way halting the process which it is controlling. In addition, it is capable of switching channels at much higher speeds than the conventional cam controller, or electromechanical timer.

A complete programme of all channels can be set up on pin board or edge switches in only a few seconds by unskilled personnel. The flexibility of this instrument being such that the programme can be set to last a total time of only a few seconds or, by slowing the master pulse frequency, can be extended to several hours, or even several days.

The Mk. III programmer shown in the photograph, has 24 modular switching channels, any or all channels can be normally open or normally closed, and can be switched at any time or several times throughout the complete programme.

The Mk. III has 9 999 increments to which any one or all the 24 channels may be set throughout the programme. If more precise resolution is required, then further module decades may be added to the counter unit to give any required number of pulses in any given time. The number of switching channels can also be increased in modular units up to at least 144 to meet the needs of any application.

**EE 89 767** *for further details*

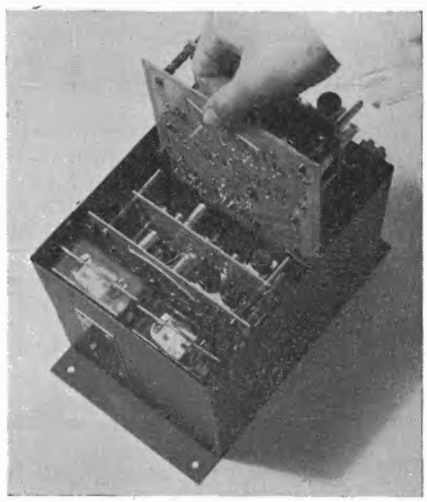
### VIBRATION MONITOR

**Dawe Instruments Ltd, Western Avenue, Acton, London, W.3**

*(Illustrated below)*

A simple device designed to monitor vibration continuously in almost any engineering field is the new Dawe type 1435 vibration monitor. It was originally designed specially for a nuclear power station contract, but a commercial version has now been developed. It consists of an electronic amplifier of printed circuit, transistorized construction. It gives a continuous electrical output proportional to the peak displacement of the structure being monitored, derived from a velocity-sensing pick-off (transducer) such as the Dawe type 1403/1 mounted on a selected point of the structure. The standard version, as supplied for the nuclear power station contract, has two separate channels contained together with power supplies and overload relays in a single compact case for wall mounting. This is a convenient arrangement enabling the channels to provide a check on each other, but single- and multi-channel units are also available with cases and mountings to customers' requirements.

Each channel provides an output signal for remote meter reading proportional to peak displacement within the frequency range 15c/s to 800c/s, more than sufficient for the majority of mechanical vibrations. Provision is made for the insertion of filters where a more restricted frequency range is required. The range

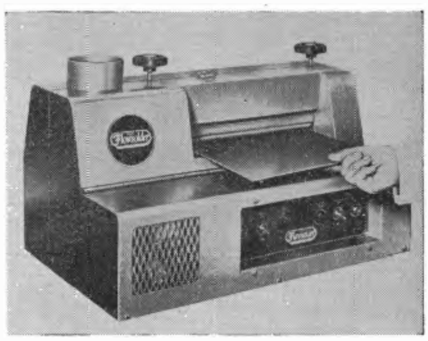


of displacement which can be monitored is 0·001in to 0·010in peak depending on the transducer used, the output signal from the monitor being 1mA maximum. Output impedance is 1k$\Omega$.

The overload relays may be preset to operate at any level from 25 per cent to 100 per cent of maximum output. In the standard version they are rated at 10A 230V a.c. and may be normally open or normally closed. They are therefore suitable for operating almost any type of warning or overload trip circuit. Separation of the relay and amplifier circuit enables the unit to continue monitoring even after an alarm has been given.

The output signal from the monitor may be measured by a meter at some remote central control point. Alternatively a local test meter is available. Under reasonable conditions, the monitor itself should require no more attention than an annual calibration check once the overload relay has been set.

**EE 89 768** *for further details*



### TINNING MACHINE

**Fry's Metal Foundries Ltd, Tandem Works, Merton Abbey, London, S.W.19**

*(Illustrated above)*

The Flowsolder Roltinner machine is designed to provide the printed circuit used with an inexpensive method of protecting circuit boards. The cleaned copper circuits are coated with Flowsolder alloy which allows longer storage life and helps in better soldering after insertion of components.

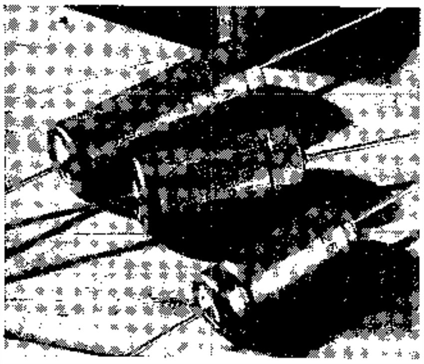
One of the main features of the Roltinner is that it takes printed circuit boards up to 13in wide of any length, and up to ⅜in thick.

The basic principle of the machine is that the circuit board is fed through two rollers: an upper and a lower roller. The upper roller is stainless steel sheathed and can be adjusted to give the correct gap between the roller and the thickness of board. This roller is spring-loaded to give constant pressure. The lower roller rotates in a bath of Flowsolder alloy.

As the circuit board passes through the two rollers, it automatically becomes coated with Flowsolder alloy at a rate of 6ft/min.

The machine is powered by a constant speed motor with a reduction gear. Power inut is 7A at 240V 50c/s operational weight is 280lb.

**EE 89 769** *for further details*



## MINIATURE ELECTROLYTIC CAPACITORS

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

*(Illustrated above)*

A new series of miniature long-life electrolytic capacitors for industrial applications is announced by Mullard Ltd. Cheaper than the C427 series, the new series, type C428 offers a greater *CV* product for a given can size and at the same time, has all the service life and reliability characteristics of the existing C427 range.

Typical service life figures at 40°C are 200 000 hours continuous operation, at an estimated failure rate of only 0·025 per cent per 1 000 hours at full rated voltage.

The use of high purity materials in the manufacture of the capacitors ensures a low leakage current. Welded connexions keep series resistance to a minimum.

The capacitance range of the C428 series covers the same values as the C427 series—2·5μF to 160μF with an extension up to 320μF. Working voltages are from 4V to 64V. The capacitors meet the IEC climatic group number 40/70/56.

**EE 89 770** *for further details*

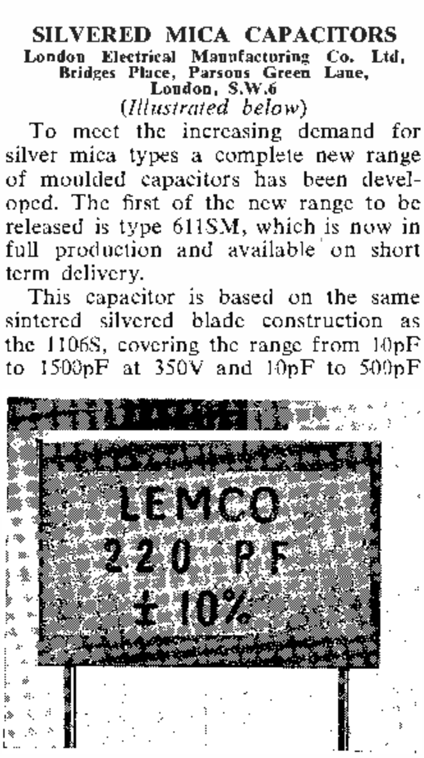
## SILVERED MICA CAPACITORS

**London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6**

*(Illustrated below)*

To meet the increasing demand for silver mica types a complete new range of moulded capacitors has been developed. The first of the new range to be released is type 611SM, which is now in full production and available on short term delivery.

This capacitor is based on the same sintered silvered blade construction as the 1106S, covering the range from 10pF to 1500pF at 350V and 10pF to 500pF




at 750V. Moulded in epoxy resin, it meets the DEF5011 humidity category H5 and measures only 15·4mm × 10·5 mm and 4·0mm, with leads pitched on 10mm centres for printed circuit board mounting.

An advanced silvering process, covered by world patents, is now used on LEMCO silver mica blades, resulting in improved life and cyclic stability. This new process has been stringently tested and shown to enhance the already excellent characteristics of the basic 1106S sintered capacitor.

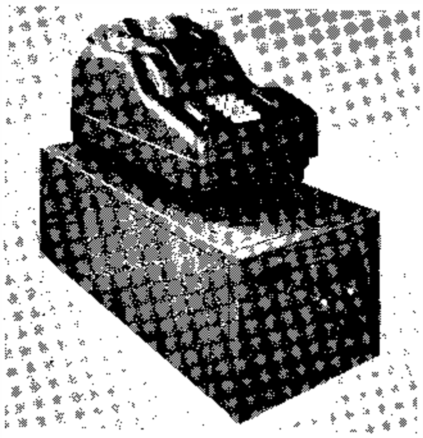
**EE 89 771** *for further details*

## ANALOGUE TO DIGITAL CONVERTOR

**Joyce, Loebl & Co. Ltd, Electron House, Princesway, Team Valley, Gateshead, 11, Co. Durham**

*(Illustrated below)*

This unit provides a simple method of automatically recording large quantities of finite information for subsequent



processing. Any succession of events represented by analogous quantities, mechanical movement, resistance, voltage, etc., can be measured, logged in series and event order and presented digitally. It can easily be added to a wide range of existing scientific instruments, spectrophotometers, pH meters, etc., to give digital print out or display facilities in most cases without modification to the original instrument.

Designed as a low price unit, it is small in size and with solid state and thin film circuits is reliable in operation. The input transducer provides an impedance of not more than 5kΩ. Where the input to the instrument is a direct voltage, the output impedance of the source should not exceed 5kΩ.

Input voltages are converted to frequencies within the range 100 to 1100c/s, with a conversion accuracy of better than 0·1 per cent. (Provision is made to off-set non-linearities to frequencies below 100c/s.) Voltages are integrated over periods of up to one second and frequencies are counted across a three-stage decade counter. Input signals of -12V initiate two-stage counters for the logging of series of runs and events within the series. Extra decades can be added to record up to a maximum of 999 series and 999 events within any series.

Zero set is adjusted from the face panel of the instrument and drift is ±1 cycle per hour, i.e. better than 1 digit, after the initial warm-up period. Incorporated is a potentiometer for scaling the instrument within the 0 to 999 range. The range setting of the convertor can be arranged to give full-scale recording for excursions within the range 0 to 999.

The output printer is a standard commercially available unit, the ADDO type No. X41E, with ten print head positions and a line print out time of less than two seconds enabling an integration period of one second per measurement. Where measured values fall outside the range and scale factor setting, an asterisk, is printed out in column one, the remaining print head positions are allocated, three digits for series reading, three for the event and three for the measured value. A version of the convertor is in preparation giving either display or punched tape outputs.

**EE 89 772** *for further details*

## STEPPING SWITCH

**Flight Refuelling Ltd, Wimborne, Dorset**

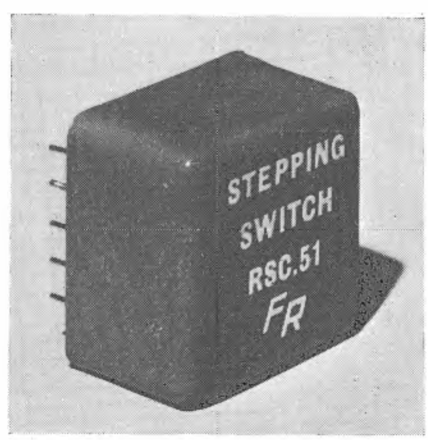
*(Illustrated below)*

Now available from the Industrial Electronics Division of Flight Refuelling Ltd, the FR type RSC-51 stepping switch offers a much improved packing density over its predecessor, the RSC-41. In stepping switch configurations using a number of RSC-51 modules each input pulse causes the stepping switch to advance one position and has an operating speed many times faster than a mechanical selector switch with a maintenance-free operating life of up to 100 million operations.

Each position is provided with two completely isolated output contacts, each a Hamlin reed switch, providing the module with all the advantages of the dry reed switch.

The RSC-51 can be driven from a low-power transistor if desired and no radiation interference is produced. Dimensions are 1·30in square by 0·83in deep.

The nominal hold current is 10·9mA, the maximum operating speed 150p/s and the minimum length of drive pulse 2msec.




**EE 89 773** *for further details*

# SHORT NEWS ITEMS

**Applications of Thin Films in Electronic Engineering** will be the subject of a Conference to be held jointly by the Institution of Electronic and Radio Engineers and the Electronics Division of the Institution of Electrical Engineers at Imperial College of Science and Technology, London, on 11 to 15 July.

The Conference will deal with the following subjects:

(a) Preparation of Thin Films
(b) Thin Films and Integrated Circuits
(c) Computer Applications

The final group of papers will deal with other applications such as, magnetoresistive films as strain gauges; thermoelectric thin film devices; luminescence in thin films; space-charge-limited currents in dielectric thin films; hybrid thin-film/semiconductor circuits.

The joint Organizing Committee will welcome offers of papers on these subjects or on related topics. Details should be sent as soon as possible accompanied by a 200-word synopsis, to the Secretary of the Joint Conference on Thin Films, c/o The I.E.R.E., 8-9 Bedford Square, London, W.C.1.

**An International Conference on Electronic Switching** is to be held at the UNESCO Conference Hall, Paris, on 28 to 31 March.

This Conference is being organized by the Union of International Technical Associations (U.A.T.I.), the Societe Française des Electroniciens et des Radio-electricians (S.F.E.R.) and sponsored by the Union International des Telecommunications (U.I.T.) and the Federation Nationale des Industries Electroniques (F.N.I.E.).

Among the subjects to be dealt with are:

(a) Connexion Networks for Spatial-Switching Systems
(b) Time Division Electronic Switching Systems

Full details on registration etc. are obtainable from the Colloque Internationale de Commutation Electronique, 16 rue de Presles, Paris, 15e.

**A Three-Day Conference** is to be held at the Congress Hall in the International Centre on 4 to 6 May during the Hannover Fair and is being organized by the Scientific Committee and the Communications Engineering Society of the VDE in conjunction with the Hannover Fair Authorities.

The first two days, will be devoted to Components and Applications, and the third day to Electronics in Aerospace.

Further information is obtainable from the Elektrotechnishe Gesellschaft Hannover e.V. im VDE 3000 Hannover, Lange-Hop-Str. 18 Germany.

**The United States Trade Centre** at Frankfurt-am-Main is to hold a special Exhibition of Production and Packaging Equipment for the Electronic Industry on 12 to 19 January at which 35 leading American Manufacturers will display their products.

Further information is obtainable from the Trade Centre whose address is: 6 Frankfurt-am-Main, Bockenheimer Landstrasse 2/4.

**The time signal service** at present provided by the Rugby v.l.f. transmitter GBR on 16kc/s will, during the first six months of this year, be transferred to Criggion v.l.f. transmitter GBZ on 19·6kc/s.

Criggion will provide the service from 0001 on 1 January, until approximately 30 June when the time signal service will revert to Rugby, but users are advised to retain their capability for reception on 19·6kc/s.

The reason for the withdrawal of Rugby is to enable the transmitters to be modernized for frequency-shift keying.

In addition to the existing on-off morse code transmissions on 16kc/s it is intended to transmit 50-baud teleprinter signals (i.e. element length 20msec) by frequency-shift keying. The lower frequency will be 15·95kc/s and the upper one 16kc/s. The present accuracy of the 16kc/s carrier will be retained and measures are being taken to prevent a phase shift in the wave when the aerial is retuned to change from on-off to frequency-shift keying and vice-versa.

**BINDING OF VOLUMES**

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

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

**The Hong Kong Royal Observatory** is to install a special radar system for locating the centres of tropical storms or typhoons within 150 to 200 miles of Hong Kong.

The radar will be sited at the Observatory's station at Tate's Cairn, a 1 890ft-high mountain north-east of Hong Kong Island. Radar information will be relayed to the airport and the Observatory by a radar link with the station. The radar will be protected by a large plastic dome specially strengthened to withstand the typhoon winds.

The extreme wind speeds and the heavy rain associated with storms require special techniques to ensure that the radar can operate in the winds and at the same time penetrate the heavy rainfall to enable the centres of distant storms to be seen.

The radar system is to be supplied by Plessey Radar Ltd and the link equipment by Pye Ltd, Cambridge.

**The 1966 Television and Radio Show** is to be held at Earls Court, London on 22 to 26 August for trade visitors only.

Participation in the exhibition will be open to all home and overseas manufacturers of television and radio sets, high fidelity audio equipment, radiograms, record players, amplifiers, disks, tape recorders, musical instruments, transmission and studio equipment and allied products and services.

Further details are obtainable from Industrial and Trade Fairs Ltd, Commonwealth House, 1-19 New Oxford Street, London, W.C.1.

**The British Standards Institution** has published a separate catalogue of the publications of four international standards bodies—ISO (International Organization for Standardization), IEC (International Electrotechnical Commission), CEE (International Commission on the Rules for the Approval of Electrical Equipment) and CISPR (International Special Committee on Radio Interference)—which expands and brings up to date to 30 June, 1965 the relevant information in the current British Standards Yearbook.

The new catalogue includes a numerical list giving the title of each international publication, a brief summary of its contents and an indication of its relationship with the equivalent British Standards to which there are corresponding international publications with the appropriate cross-references to these publications.

Free copies may be obtained by members from the BSI Sales Branch, 2 Park Street, London, W.1. Non-members can buy copies at the full price of 7/6 (Postage 9d will be charged extra).

**The Institution of Electronic and Radio Engineers** is organizing what is believed to be the first Conference to be held in Europe on the applications of electronic engineering to oceanography. The Conference will be held at the University of Southampton on 19 to 22 September.

A feature of the Conference will be the opportunity which it will provide for oceanographers to state their problems as users and for electronic engineers to put forward means of solution, and papers will therefore be welcomed from both points of view.

Further information and offers of papers (which should preferably be accompanied by a 200 word synopsis) should be sent to the Secretary of the Conference on Electronic Engineering in Oceanography, The I.E.R.E., 8-9 Bedford Square, London, W.C.1.

**Pye-Ling Ltd** has received an order valued at £200 000 from the European Space Research Organization for the supply of vibration equipment for the ESRO experimental test centre at Noordwijk in Holland. The test centre, when completed this year, will be the largest and best equipped space research laboratory outside the United States.

The vibrators will be used to test models and full-sized prototypes of rockets and space probes by simulating the conditions experienced when in space.

Three systems which include power amplifiers, control and monitoring facilities will provide thrust capabilities of 4 000, 8 000 and 15 000lb.

**Decca Radar Ltd** has now received its 20 000th order for marine radar. This order which came from France is for the new Transar Series Type RM314 relative motion radar for installation in a new stern trawler under construction in France.

The total has been reached towards the end of a record year for Decca marine radar sales. In January 1965 a total of 17 000 orders was passed; in May 18 000 orders; in August, 19 000 orders. By the end of 1965 a total exceeding 3 500 orders will have been received in 12 months.

It is estimated that approximately half the world's radar fitted merchant ships of over 100 g.r.t. are using Decca marine radar. Decca naval radars are also in service with 48 of the world's navies.

**Exports of valves,** tubes and semiconductor devices during the third quarter of 1965 reached a total of £3 165 115 according to figures issued by VASCA and BVA (Electronic Valve and Semi-Conductor Manufacturers' Association and Brtiish Radio Valve Manufacturers' Association) based on the Customs and Excise returns.

The totals for the nine months to 30 September, 1965, show that the exports of semiconductor devices have increased by 46 per cent over the corresponding period of 1964, and the exports of valves and tubes have increased by 11 per cent. (Export figures for the first nine months of 1964 were—semiconductor devices: £1 369 956; valves and tubes: £6 636 268.)

**The Radar Division** of Cossor Electronics Ltd has received an initial order valued at £175 000 for airborne transponders, from the Ministry of Aviation, to be fitted in the McDonnell Phantom F4M aircraft ordered for use by the Royal Air Force.

These transponders were developed to Ministry of Aviation high performance aircraft specifications and are designed and engineered to meet or exceed the most stringent requirements for this type of equipment.

The unique design provides compatible operation with all known military and civil s.s.r./i.f.f. systems, either in use or projected, and the operational parameters make the equipment suitable for installation in a very wide range of military and civil machines.

**The Ministry of Aviation** and the Electronic Engineering Association have completed arrangements for the incorporation of British electronic equipment in the Hercules C.130-K aircraft which are to be purchased from the United States.

Following the announcement by the Government of cancellations in the programme for British manufactured aircraft for the R.A.F., the Electronic Engineering Association sought to discuss with the Minister of Aviation the means by which the maximum possible content of British electronic equipment could be incorporated in the Hercules and Phantom aircraft to be purchased from the U.S.A., having regard to the essential need to meet the required early delivery dates.

In the case of the Hercules, it was proposed that more modern solid-state equipment, already in service with the R.A.F., should be used in place of the U.S. equipment hitherto fitted in the United States Air Force C.130-K (Hercules) aircraft.

The following is a list of British electronic equipment manufacturers and the equipment they will supply:

Ekco Electronics Ltd
- Weather Radar

The Plessey Co. Ltd
- U.H.F./V.H.F. Communications

The Marconi Co. Ltd
- V.H.F. Communications
- V.O.R./I.L.S.
- A.D.F.

Ultra Electronics Ltd
- Cabin Address System
- Intercom System

Standard Telephones & Cables Ltd
- Radio Altimeter

The Decca Navigator Co. Ltd
- Navigating Computing & Display System comprising:
- Doppler Sensor
- Navigation Computer
- Lat/Long Display
- Along/Across Track Display
- Auxiliary Across Track Display
- Roller Map Display
- Decca Navigator System

S. Smiths & Sons Ltd
- Autopilot

Cossor Electronics Ltd
- I.F.F.

**The Marconi Co. Ltd** is to instal an educational closed circuit television system at Edinburgh University consisting of nine fully automatic television cameras and fifty-one receivers.

The nine cameras and their associated sound channels will be connected to a central Control Room, from which up to three different programmes can be distributed simultaneously to selected viewing rooms throughout the building.

The cameras used in this system will be the new Marconi V322 series, designed specifically for educational television. Each camera is fully automatic in operation, being able to cater for a full range of lighting conditions without attention from an operator. In normal use, the ON/OFF switch is the only electronic control that need be touched.

Four of the cameras will be mounted on lightweight tripods for mobile use. These cameras will be the studio version type V322B fitted with integral electronic viewfinder.

The other five cameras type V322A will be permanently mounted in the lecturers bench in selected lecture theatres. These cameras do not incorporate the viewfinder. Each camera will be mounted behind a screen and will be used either as electronic blackboards, or for document transmission.

**The Royal Air Force Transport Command** has placed an order with Hawker Siddeley Dynamics for two T.R.A.C.E. automatic test equipments for general purpose testing of electronic equipment on the new VC-10 and Belfast aircraft of Transport Command.

Punched tape programmes which dictate the tasks to be performed by T.R.A.C.E. (Tape-controlled Recording

Automatic Test Equipment), and coupling units to connect the equipment to the item under test, will be prepared initially for thirty-nine different aircraft components and sub-assemblies.

The Royal Air Force T.R.A.C.E. will closely resemble the equipment supplied to British Overseas Airways Corporation at London Airport, and the Admiralty at Portsmouth Dockyard.

**The British Electrical and Allied Manufacturers' Association** (BEAMA) has published a specification guide for electronically controlled adjustable speed drive equipment (Publication No. 208, price 5s.)

This publication has been submitted to the British Standards Institution as a basis for a proposed British Standard.

The next specification guide to be published by BEAMA will be one on photo-electric switching units.

**The Sectional Meeting of the World Power Conference** is to be held at Tokyo on 16 to 20 October and the theme of the Conference will be 'Problems of Future Years in Energy Utilization.'

Sixteen papers are being presented by the British National Committee.

Further details are obtainable from The Secretary, British National Committee, World Power Conference, 201-2 Grand Buildings, Trafalgar Square, London, W.C.2.

**The Bromley Technical College** has installed in its Engineering Department a Mullard Electronics Trainer to assist its lecturers to prepare students for the City and Guilds of London Institute Examinations in electronics and telecommunications.

The Trainer has been supplied with 21 circuit panels which make possible the construction of many circuits including a complete superhet receiver, and a transmitter. There is also a panel specifically designed for measuring valve characteristics.

**A symposium on Collection and Processing of Field Data** will be held in Canberra (Australia) from 30 August to 2 September 1966, under the auspices of the Commonwealth Scientific and Industrial Research Organization. The symposium has been accepted as an Australian contribution by both the International Hydrological Decade, and the International Biological Programme.

The meeting will include discussion of the instrumental, sampling and data treatment problems peculiar to measurement in the field environment, and contributions by electronics and systems engineers on recent technological developments relevant to these problems.

It is intended that the symposium foster collaboration in this direction between the engineering sciences, and field workers in micrometeorology, hydrology, ecology, oceanography, animal physiology and allied disciplines.

Further information can be obtained from Dr. E. F. Bradley, Division of Plant Industry, P.O. Box 109, Canberra City, A.C.T. Australia.

**Dawe Instruments Ltd** has introduced two new divisions within the Company, namely the Industrial Division and the Instrument Division, in addition to the Power Ultrasonic and Non-Destructive Testing Divisions introduced last year.

The Industrial Division will develop and market sound and vibration measurement and analysis equipment, stroboscopes and the digital torquemeter originally developed in association with the Ford Motor Co. Ltd.

The Instrument Division will be responsible for all laboratory test equipment. Its own range of equipment will be supplemented by the Danbridge Decade Resistance Capacitance and Inductance Boxes now marketed in the U.K. by the company.

# 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: 6 January.
Lecture: Lasers and Associated Devices.
By: G. G. Macfarlane.

**I.E.E./R.Ae.S. London Joint Group on the Applications of Electricity and Electronics in Aircraft**

Date: 25 January. Time: 6 p.m.
Discussion: The Impact of Microminiaturization on Aircraft Systems.
Opened by: A. A. Shepherd.
Held at: The Royal Aeronautical Society, 4 Hamilton Place, W.1.

### Control and Automation Division

Date: 4 January.
Lecture: Satellite Control.
By: E. G. C. Burt.
Date: 11 January.
Discussion: Alpha-Numeric Cathode-Ray-Tube Displays.
Opened by: H. C. A. Hankins, A. T. Starr and L. W. Whittaker.
Date: 18 January.
Colloquium: Human Operator Studies.
(All wishing to attend must apply to the Secretary for a ticket. No fee for IEE members; non-members £1 1s.)
Date: 25 January.
Discussion: The Use of Programmed Learning for Remedial Work in Higher Education.
Opened by: D. Unwin and A. K. Spencer.
Date: 27 January.
Lecture: Ionization Techniques Applied to Gas-Flow Measurement.

### Electronics Division

Date: 4 January.
Lecture: Naturally Occurring Thermal Radiation in the Range 1-10Gc/s.
By: D. L. Croom.
Lecture: Aerial Noise Temperature at 5 650Mc/s.
By: E. Denison and G. L. Rogers.
Date: 12-14 January.
Conference: V.H.F. and U.H.F. Mobile Communication Systems and Equipment.
(All wishing to attend must register; forms available from the Secretary.)
Date: 17 January. Time: 2.30 p.m. and 5 p.m.
Colloquium: Active Filters.
(All wishing to attend must apply to the Secretary for a ticket. No fee for IEE members; non-members £1 1s.)
Date: 19 January. Time: 6 p.m.
Lecture: Industrial Applications of Electroluminescence.
By: D. Reaney.
Date: 24 January.
Lecture: Recent Studies of Dispenser Cathodes.
By: H. Ahmed, C. E. Maloney, J. D. Sankey and A. H. Beck.
Date: 25 January.
Lecture: Method for the Prediction of the Fading Performance of a Multisection Microwave Link.
By: K. W. Pearson.
Date: 27 January.
Lecture: Optical Surface Waves.
By: C. Kao.
Date: 28 January. Time: 2 p.m.
Colloquium: Microwave Noise-Measuring Techniques.
(All wishing to attend must apply to the Secretary. No fee for IEE members; non-members £1 1s.)
Date: 31 January.
Lecture: The Definition and Properties of the Delta Function.
By: R. F. Hoskins.
Date: 31 January.
Colloquium: The Problems of Radiation Pressure and Associated Mechanical Forces.
(All wishing to attend must apply to the Secretary. No fee for IEE members; non-members £1 1s.)

### Power Division

Date: 5 January.
Discussion: Operating Experience with Deck Machinery.
Opened by: W. S. C. Jenks and D. Campbell.
Date: 7 January.
Discussion: Tracking Insulation.
Opened by: A. G. Day, P. J. Jackson and A. W. Stannett.
Date: 20 January.
Lecture: Electric Process Heating for the U.K.A.E.A. Works.
By: E. R. Davies, R. J. Gamble and J. B. Silleto.
Date: 24 January.
Colloquium: Gaseous Insulation.
(All wishing to attend must apply to the Secretary for a ticket. No fee for IEE members; non-members £1 1s.)
Date: 26 January.
Lecture: The 250kV D.C. Submarine-Power-Cable Interconnection Between the North and South Islands of New Zealand.
By: A. L. Williams, E. L. Davey and J. N. Gibson.

## THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meetings will be held in the I.E.R.E. Lecture Room, 9 Bedford Square, London, W.C.1.

### Radar Group

Date: 19 January. Time: 6 p.m.
Lecture: Improved Radar Visibility of Small Targets in Sea Clutter (through decorrelation by a Rapidly Rotating Antenna).
By: J. Croney.

### Computer Group

Date: 26 January.
Paper on: Linking Computers.
(Joint I.E.R.E./I.E.E.)

## THE TELEVISION SOCIETY

All meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3. and commence at 7 p.m.

Date: 13 January.
Lecture: Problems Connected with the use of Colour Film for Colour Television.
By: F. P. Gloyns.
Date: 28 January.
Lecture: U.H.F. Translators.
By: W. J. Morcombe.

# NOUVELLES Réalisations

*Traduction des pages 48 à 53*

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

### MICROVOLTMÈTRE À FRÉQUENCE ACOUSTIQUE

**Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire**

*(Illustration à la page 48)*

Le microvoltmètre BF, type 705, vient d'être ajouté à la série des instruments à fiches du système Dymar. Il s'agit d'un voltmètre à tube conçu pour mesurer les tensions alternatives dans la gamme des fréquences acoustiques bien que sa réponse de fréquence s'étende jusqu'à 300 kHz avec une précision totale.

Cet instrument se caractérise par le fait qu'il a une "entrée équilibrée" qui permet la mesure directe des tensions push-pull par rapport à la terre.

Le rejet "en phase" sur la plus grande partie de la gamme de fréquence est supérieur à 60 dB, ce qui autorise la mesure de tensions équilibrées en dépit de tensions d'interférence nettement plus élevées. L'une ou l'autre des entrées peut être utilisée pour mesurer des tensions déséquilibrées par rapport à la terre.

La sensibilité d'entrée est de 100 $\mu$V pour une déviation sur la totalité de l'échelle avec un niveau de bruit inhérent inférieur à 20 $\mu$V. Ce niveau de bruit peut être réduit à 8 $\mu$V en limitant la largeur de bande à 10 kHz.

L'impédance d'entrée sur chaque entrée est de 10 M$\Omega$ et les bornes de sortie dont est muni l'appareil permettent de l'utiliser comme préamplificateur sensible pour actionner des oscilloscopes ou des amplificateurs de puissance.

**EE 89 751** *pour plus amples renseignements*

### AMPLIFICATEUR À LARGE BANDE

**Telequipment Ltd, 313 Chase Road, Southgate, London, N.14**

*(Illustration à la page 48)*

La société Telequipment Ltd a créé un nouvel élément dans la gamme des amplificateurs à fiches. Il s'agit du type H, conçu pour l'amplification à large bande et particulièrement indiqué pour la mesure du temps de montée.

L'appareil comporte deux prises d'entrée avec un isolement de 60 dB entre les entrées qui s'obtiennent par commutateur. Les largeurs de bande vont de la tension continue à 25 MHz (−3 dB) avec une sensibilité de 100 mV/cm à 50 V/cm, et de la tension continue à 5 MHz (−3 dB) avec une sensibilité de 10 mV/cm. La sensibilité dans chacun des cas est branchée dans un rapport de 1:2:5, et une commande de gain assure la variation continue entre les gammes d'atténuation à degrés fixes.

L'amplificateur type H, de même que les autres cinq amplificateurs à fiches produits par Telequipment Ltd, est pleinement interchangeable avec les oscilloscopes Telequipment type S43, D43 et D43R, ainsi qu'avec le nouvel oscilloscope D53 qui vient d'être sorti par cette société.

**EE 89 752** *pour plus amples renseignements*

### CONTRÔLEUR DE FINISSAGE DE SURFACE

**Payne Products International Ltd, Buckingham Avenue, Trading Estate, Slough, Buckinghamshire**

*(Illustration à la page 48)*

La société Payne Products International Ltd, productrice des machines de finissage Lapmaster, vient d'annoncer la mise au point d'un nouveau microcontrôleur, le Diavite Mark II, pouvant mesurer toutes les gammes connues de finissage de surface. Fonctionnant sur courant alternatif monophasé de 110 ou de 220 V, le nouvel appareil peut mesurer des moyennes d'entre-axes de 0 à 300 $\mu$ pouces. Cette échelle est divisée au moyen d'un bouton-poussoir à quatre sections, allant jusqu'à 10, 30, 100 et 300 $\mu$ pouces.

L'instrument comporte deux distances de coupure: 0,003 pouce pour les sections concaves ou convexes très fines ou 0,03 pouce pour les sections à profil plat.

L'appareil comporte une lumière de tolérance qui peut être réglée aux valeurs de limite prévues pour tout composant devant être soumis à un contrôle. Si la limite est dépassée, un voyant avertisseur rouge s'allume sur la surface.

Il comprend également trois styles mesurant 0,4 pouce, 0,8 pouce et 0,16 pouce. La vitesse de déplacement du style est de 0,05 pouce/sec. Le nouvel appareil est portatif et ne pèse que 6 kg.

**EE 89 753** *pour plus amples renseignements*

### MODULES DE COMPTAGE DE DÉCADES

**Quarndon Electronics Ltd, Slack Lane, Derby**

*(Illustration à la page 48)*

Cette gamme de modules compteurs de décades a été conçue pour les systèmes industriels ainsi que pour les laboratoires qui ne sauraient justifier les frais d'acquisition d'un compteur électronique complet. Ces éléments modulaires constituent un système d'une grande souplesse d'emploi pour les applications spécialisées.

Ils occupent une surface de panneau frontal de 12,06 cm × 3,81 cm et leur profondeur est de 13,97 cm, des terminaisons remplissant la plaquette de circuit imprimé standard à 32 directions. Le comptage est affiché sur un tube indicateur numérique de type classique.

Les modules sont fournis soit avec des dispositifs semi-conducteurs au silicium soit avec des dispositifs au germanium. La vitesse de comptage maxima du module DCM 501 est de 1 MHz. D'autres modules au germanium ont des vitesses de comptage de 5 MHz et de 12 MHz. Les modules au silicium ont des vitesses de comptage de 200 kHz, 2 MHz et 30 MHz. Il existe également des compteurs réversibles. La décimale est indiquée par une ampoule au néon miniature, actionée de l'extérieur.

L'impulsion d'entrée exigée dépend dans une certaine mesure de son temps de montée. Pour la vitesse maxima, une impulsion de 2,5 V avec un temps de montée de 500 nsec est nécessaire.

La sortie de l'élément suffit pour actionner un autre élément, de sorte qu'on peut ainsi réaliser un compteur multinumérique. Il existe des sorties décimale codées binaires de 8-4-2-1 pour l'affichage à distance.

Le module peut être réenclenché à zéro en appliquant une impulsion de réenclenchement à la borne de réenclenchement. Il existe en outre des étalons de fréquence au cristal, des éléments de porte et des éléments d'affichage à distance qui complètent la gamme et permettent de réaliser des compteurs complets.

**EE 89 754** *pour plus amples renseignements*

## KLYSTRONS RÉFLEX

**The English Electric Valve Co. Ltd, Chelmsford, Essex**

(*Illustration à la page 49*)

La société English Electric Valve Co. Ltd a réalisé une nouvelle série de klystrons réflex, bande "X", pour répondre à la demande croissante d'oscillateurs locaux robustes et peu onéreux ayant, en outre, une gamme étendue de coefficients de fréquence/température.

Ce dernier groupe de tubes a été mis au point suivant les mêmes méthodes sévères de production que celles employées dans la construction des robustes klystrons EEV pour usage militaire. Les avantages évidents qui découlent de l'expérience acquise grâce à ces derniers se manifestent dans le rendement et la fiabilité des nouveaux tubes.

Cette série de klystrons se caractérise par le fait qu'on peut en choisir un dont le coefficient de fréquence/température convient à l'application particulière à laquelle il est destiné. Le modèle K3007, par exemple, a un coefficient de $-130$ kHz/°C, c'est à dire un coefficient égal à celui de la plupart des magnétrons de la bande "X" vendus dans le commerce. En plus, les nouveaux tubes offrent un degré étendu d'interchangeabilité, permettant la permutation de paramètres de rendement pour répondre aux besoins individuels dans le cas des klystrons non-standard.

Parmis les tubes qui composent cette série, il faut signaler le K3003, qui est identique au point de vue électrique au K3020, sa variante à câble volant. Le K3003 est en effet particulièrement indiqué pour l'usage marin, ses caractéristiques excellentes de dérive de fréquence lui ayant permis de demeurer en deça d'une variation de 10 MHz au cours d'un contrôle de plus de 3000 heures. Ajoutons encore que la modulation de fréquence de pointe du tube se maintient, suivant un maximum de spécification de 200 kHz sous des charges vibratoires de 10 g.

**EE 89 755** *pour plus amples renseignements*

## INSTRUCTEUR MICROLOGIQUE

**Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire**

(*Illustration à la page 49*)

L'instructeur micrologique LD-20 dérive du modèle bien connu LD-15 et emploie pour la logique des circuits intégrés à semi-conducteurs. L'emploi de ces modules constitue en matière de construction électronique une question d'interconnexion entre les circuits plutôt que la réalisation détaillée de circuits individuels. L'instructeur micrologique offre à l'utilisateur de circuits intégrés semi-conducteurs la possibilité d'acquérir une expérience pratique de tous les aspects des travaux avec les circuits numériques intégrés. Chaque circuit intégré à semi-conducteurs se compose d'un circuit complet numérique ou linéaire dans lequel tous les composants sont fabriqués à partir d'une petite tranche semi-conductrice, enfermée dans une capsule à transistors (telle que celle de la dimension TO-5) ou dans un boîtier plat, dont les dimensions-type sont de 6,6 mm$^2$ × 1,3 mm d'épaisseur.

Les éléments à circuit intégré semi-conducteur de l'instructeur LD-20 fonctionnent à partir d'une ligne d'alimentation de 3,6 V seulement, à un courant inférieur à 1 mA, ce qui indique bien la tendance à la logique à faible niveau. Ces niveaux de faible puissance sont évidemment un grand avantage dans la réalisation d'appareils compacts.

L'instructeur LD-20 comprend 15 portes de circuit intégré semi-conducteur, avec cinq amplificateurs intermédiaires à deux étages supplémentaires pour actionner les lampes indicatrices. Le LD-20 est également d'une plus grande souplesse d'emploi que le LD-15 puisque cinq tumblers à deux directions sont prévus pour l'entrée, en plus du cadran téléphonique. Ces tumblers donnent des tensions négatives ou positives facultatives aux douilles adjacentes qui s'adaptent à la logique. Bien que les circuits logiques constituent fondamentalement des portes NI (dans lesquelles OV = 0 et la tension d'alimentation positive = 1), en inversant les conventions de signe, ces quinze portes peuvent être utilisées pour la fonction NI-ET en actionnant simplement les tumblers d'entrée.

L'instructeur micrologique comprend un bloc d'alimentation fonctionnant sur secteur et donnant donc les tensions continues correctes; il est fourni avec le manuel d'instruction numérique LANDEC de plus de 100 pages, ainsi qu'avec le manuel micrologique (établi par S.G.S.—Fairchild) et un manuel spécialement rédigé pour l'instructeur micrologique LD-20.

**EE 89 756** *pour plus amples renseignements*

## BLOC D'ALIMENTATION STABILISÉE T.H.T.

**Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex**

(*Illustration à la page 49*)

Le bloc d'alimentation stabilisée PMD 101 de 5 kV est l'un des premiers éléments produits par la Division d'instruments de précision de la société Cathodeon Electronic Ltd qui vient d'être formée.

Le nouveau bloc donne des sorties positives ou négatives par rapport à la masse à n'importe quel courant préréglé jusqu'à la pleine puissance nominale de 5 mA. La commande de préréglage se trouve sur le panneau frontal.

Le disjoncteur à maxima est un module transistorisé d'un dessin spécial et qui peut être réglé pour répondre dans la gamme de courant de 2 à 5 mA. Le courant de court-circuit n'est que de 20% supérieur au réglage de seuil de surcharge.

La régulation du bloc d'alimentation est supérieure à 0,1% sur toute la gamme de tension et l'instrument de mesure à double échelle peut être utilisé pour indiquer le courant ou la tension.

Le PMD 101 trouve des applications dans des domaines tels que l'alimentation des multiplicateurs photoélectriques et l'alimentation T.H.T. des tubes cathodiques.

Le PMD 101 est fourni dans un coffret pour l'emploi sur banc d'essai et il est également livrable pour montage sur bâti standard de 48,2 cm.

**EE 89 757** *pour plus amples renseignements*

## CONTRÔLEUR DE TEMPÉRATURE

**C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5**

(*Illustration à la page 50*)

La société C.N.S. Instruments Ltd vient de mettre au point un nouveau contrôleur de température, le S.R.2, qui est un appareil de haute précision et entièrement transistorisé employant des réacteurs saturables pour le contrôle proportionnel de température.

Cet appareil comprend un certain nombre de nouvelles caractéristiques dont un dispositif de limitation du courant (correspondant à un rapport de puissance de 30:1) pour le contrôle des fours au platine et au molybdène.

Un commutateur électrotechnique permet de mesurer soit la proportion de sortie de puissance du contrôleur soit de lire la tension déséquilibrée de pont à thermomètre de résistance au platine et, partant, d'évaluer la déviation de température à partir du point de réglage. Les commandes comportent une surmodulation extrêmement réduite pour permettre les essais répétés à une température donnée.

Grâce à l'emploi de circuits imprimés, on peut remplacer rapidement la plus grande partie de l'ensemble du circuit en cas de panne de composants.

L'appareil peut être utilisé avec des charges résistives ou inductives et le réacteur saturable qui contrôle la puissance du four peut résister à des phénomènes transitoires de tension excessive et à des surcharges de courant.

Le S.R.2 comprend un potentiomètre C.N.S. de précision à 50 tours et à curseur qui peut être réenclenché à une partie dans 5000, permettant ainsi un réglage de température extrêmement précis.

Les essais effectués sur des installations typiques de petits fourneaux ont montré que, dans des conditions normales, la stabilité du contrôle à partir d'une alimentation secteur non-stabilisée est supérieure à ±0,02% à 1000°C. Lorsqu'on emploie une thermomètre à résistance au platine C.N.S., l'appareil comprend, en outre, la compensation automatique de la tension secteur qui répond, à quelques cycles près, aux variations de la tension secteur.

La dissipation de puissance réduite (40 VA) des contrôleurs permet un entassement serré sans surchauffe excessive. Le refroidissement forcé n'est donc pas nécessaire.

**EE 89 758** *pour plus amples renseignements*

### VOLTMÈTRE NUMÉRIQUE

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

*(Illustration à la page 50)*

L'instrument Dynamco DM2006 est un voltmètre numérique constitué de corps solides et utilisant une nouvelle méthode de conversion de tension qui assure une économie remarquable de composants de circuit. La société Dynamco (précédemment la Digital Measurements Ltd) a pu produire ainsi un voltmètre d'un prix modique sans sacrifier la qualité du rendement et les normes de construction.

Cet instrument se compose d'un élément d'affichage et de mesure de base (échelle de 9 999) et d'un module interchangeable qui détermine la gamme voulue. L'élément interchangeable le plus simple comporte quatre gammes de 100 $\mu$V à 1 kV et donne une précision de lecture de 0,05%. On peut obtenir des signaux de lecture qui actionnent les dispositifs de sortie en utilisant des cartes interchangeables à fiche et qui ne portent aucun préjudice au rejet de mode commun de l'instrument. La société Dynamco Instruments Ltd projete la réalisation de modules d'une sensibilité plus élevée et assurant le repérage automatique, la mesure de courant alternatif et la mesure de résistance. Ces instruments seront livrables au fur et à mesure du développement des programmes de mise au point.

Les constructeurs ont attaché une importance considérable à la fiabilité. Le DM2006 a, ainsi, été construit par l'emploi de cartes de circuit imprimé en époxyde de verre, de transistors au silicium et de composants du type militaire.

**EE 89 759** *pour plus amples renseignements*

### FILTRE PASSE-BAS JUMELÉ

**The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey**

*(Illustration à la page 50)*

Le nouveau filtre basse fréquence de la société Wayne Kerr se compose de deux canaux identiques pouvant être utilisés indépendamment soit en cascade soit en passe-bande. Une commande d'accord unique fonctionne sur 28 gammes pour permettre le réglage précis des deux sections à n'importe quelle fréquence voulue entre 0,1592 Hz et 1592 Hz.

En tant que filtre passe-bande jumelé, le SA500 est particulièrement indiqué pour mesurer la performance de systèmes de contrôle. L'un des filtres est introduit dans le câble de sortie du système pour améliorer le rapport signal/bruit. L'autre filtre est introduit dans le parcours du signal de référence pour présenter une caractéristique identique et laisse ainsi la fonction générale de transfert du système insensible au filtre.

Lorsque deux filtres passe-bas se trouvent en cascade, le taux d'atténuation est de 36dB/octave. Dans l'état de passe-bande, on obtient un gain unitaire et un déphasage nul à la fréquence centrale et une atténuation de 18dB/octave des deux côtés.

L'impédance d'entrée de chaque canal est de 2 MΩ (résistive) et l'impédance de sortie est inférieure à 60 Ω. Le SA500 peut être fourni soit pour montage sur bâti de 48,26 cm soit dans un coffret portatif.

**EE 89 760** *pour plus amples renseignements*

### APPAREIL DE MESURE DE LA DISTORSION

**Standard Telephones & Cables Ltd, S.T.C. House, 190 Strand, London, W.C.2**

*(Illustration à la page 50)*

Conçu pour effecteur la mesure du balayage dynamique de systèmes microondes et d'autres matériels de télécommunication, le distorsiomètre 74834-A mesure l'amplitude et la distorsion de phase du signal transmis causées par la non-linéarité dans le système. Utilisé en liaison avec un oscilloscope approprié, il mesure les amplitudes différentielles de 0,05% à 100% et la phase différentielle de 0,02° à 90°.

Quatre signaux de contrôle h.f. différents sont prévus: l'un de ces signaux est à la fréquence sub-porteuse de télévision en couleur de 4,4297 MHz et les trois autres à 304 kHz, 2 MHz et 7 MHz. D'autres fréquences peuvent être prévues pour répondre à des besoins particuliers. N'importe lequel des signaux de contrôle h.f. peut être ajouté au signal de balayage basse-fréquence, et l'amplitude de ce dernier peut être préréglée de manière à ce que le signal composé puisse balayer à son tour diverses gammes d'amplitude de régime du système. La fréquence de balayage est de 81 Hz. Des marqueurs sont incorporés à l'affichage de l'oscilloscope pour faciliter la lecture de l'amplitude et de la distorsion de phase.

L'appareil peut être utilisé isolément pour la mesure de circuit, ou avec un autre appareil analogue pour la mesure d'une extrémité à l'autre; en variante le récepteur peut être utilisé à une extrémité et le générateur peut être retiré du coffret et utilisé à l'autre extrémité en liaison avec un élément de contrôle supplémentaire fourni comme accessoire.

**EE 89 761** *pour plus amples renseignements*

### ELÉMENT PHOTOÉLECTRIQUE "RÉFLEX"

**Lancashire Dynamo Electronic Products, Rugeley, Staffordshire**

*(Illustration à la page 50)*

La société Lancashire Dynamo Electronic Products (Group MI) vient d'étendre sa gamme courante d'appareils photoélectriques par l'introduction d'un nouvel élément de tête, série RPH6, pouvant être utilisé avec son appareil de contrôle transistorisé.

Le RPH 6, qui fonctionne selon le principe réflex, se compose d'une seule tête photoélectrique à distance logeant les éléments de projection et de réception. Ces composants sont disposés de manière à ce que la cellule réceptrice puisse "voir" le faisceau lumineux qui est réfléchi par la cible. L'appareil convient pour toutes les applications comportant le positionnement, le comptage et la protection contre la sur-course, ainsi que d'autres utilisations analogues.

L'élément de tête est un moulage en alliage léger et comprend deux composants principaux, le corps et le couvercle. Le couvercle amovible facilite l'accès à l'intérieur sans déranger les composants optiques. Un joint en caoutchouc assure le scellement contre les intempéries et la construction générale permet d'effectuer un montage rigide et d'éliminer ainsi un dérangement de l'alignement dû au chocs et aux vibrations.

Le montage réflex consiste en une source lumineuse standard de 8 W à laquelle on a ajouté un réflecteur de verre de 3,81 cm de diamètre × 3,81 cm de concentration et un porte-cellule.

Le réflecteur est monté sur un support en aluminium dont le réglage peut s'effectuer sur plan vertical, plan horizontal et plan inversé, afin d'assurer l'acuité de l'image de filament d'environ 50,8 cm à la limite de fonctionnement maxima.

Pour les travaux à portée réduite ou pour les conditions dans lesquelles on exige une image à filament plus petite, ou définie avec plus de précision, on peut ajouter une lentille supplémentaire d'un diamètre de 3,81 cm.

Par l'emploi de divers composants électriques et optiques pouvant être assemblés en une variété de combinaisons répondant à des conditions particulières, on peut utiliser le RPH 6 sur des distances allant de 2,5 cm à environ 6 mètres.

L'élément de base est muni d'une cellule photoélectrique semi-conductrice ORP.60, dont la résistance varie lorsqu'elle est exposée à la lumière. Le fonctionnement est satisfaisant dans des températures ambiantes allant jusqu'à 45° C.

**EE 89 762** *pour plus amples renseignements*

### GÉNÉRATEUR D'IMPULSIONS

**K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4**

*(Illustration à la page 51)*

Le générateur d'impulsions, modèle T-15, est un instrument entièrement constitué de corps solides et qui a été conçu de manière à en faire un appareil extrêmement compact utilisant des plaquettes de circuit imprimé et des composants de haute qualité. Bien que compact, cet instrument est d'une grande souplesse d'emploi. Il fournit des impulsions simples ou doubles avec polarité positive et négative simultanée. Des largeurs d'impulsions et des retards de 30 nsec à 30 msec peuvent être obtenus et l'oscillateur interne a une gamme de fréquence de 15 Hz à 15 MHz. On peut obtenir, en outre, des largeurs d'impulsion allant jusqu'a près de 100% du cycle de régime. Des impulsions positives et négatives ayant une amplitude atteignant 10 V de pointe dans 50 Ω peuvent être obtenues grâce à la commande continue d'amplitude.

L'appareil peut être déclenché par son propre oscillateur interne, par un bouton-poussoir interne à détente unique ou par une forme d'onde extérieure. On peut également obtenir la fréquence de balayage et de porte par des signaux extérieurs.

**EE 89 763** *pour plus amples renseignements*

### ELÉMENTS À RETARD DE RELAIS

**Distributeurs: D. Robinson & Co. Ltd, Victoria House, 44 Park Street, Camberley, Surrey**

*(Illustration à la page 51)*

La société Clifford & Snell Ltd vient d'annoncer la création de deux nouveaux éléments à retard électronique pouvant être employés avec le système à relais à fiches D2600.

Le premier de ces éléments est une variante du "Superslug-A," fonctionnant sur tension alternative de 110 V. Comme on peut le voir sur notre gravure, il est extérieurement identique à un relais D2600 et peut être introduit dans le même type de prise.

Il s'appelle "Superslug" parce qu'il retarde ou ralentit le déclenchement d'un relais connexe de la même manière qu'un anneau en cuivre incorporé, mais il le fait pendant une période de temps beaucoup plus longue.

Il utilise un tube à cathode froide alimenté à partir d'un circuit doubleur de tension et fournit des retards de 1 à 72 sec avec une précision de répétition de l'ordre de ±1%. Il est d'une grande fiabilité et il est employé dans le type d'applications où les relais Clifford & Snell sont utilisés c'est à dire sur les navires, les centrales, etc. Il comporte un fusible interne de précaution ainsi qu'une résistance de limitation. Le second élément est le "Superslug-B2," qui fonctionne soit sur 24 volts c.c. soit sur 50 volts c.c., par simple modification d'une connexion. Il remplace le "Superslug-B," étant d'un modèle plus perfectionné, et son prix est trois fois moindre.

**EE 89 764** *pour plus amples renseignements*

### CONTRÔLEUR DE DONNÉES

**The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex**

*(Illustration à la page 51)*

Ce nouveau contrôleur, type L.675 a été conçu par la division des télécommunications de la société M.E.L. Equipment Co. Ltd, pour répondre à la demande d'un instrument portatif pouvant mesurer la performance de matériel et de circuits de transmission de données. Il a été étudié selon la spécification de l'administration des postes britanniques pour le contrôleur 1 A Datel.

Il comprend des sections séparées de transmission et de réception pour permettre d'effectuer la transmission le long du canal d'un circuit à deux directions, cependant que la réception sur l'autre canal s'effectue à une vitesse différente.

Le nouveau contrôleur de données se distingue par son grand distorsiomètre à lecture directe et sans ambiguité pour mesurer la distorsion de polarisation moyenne et la valeur de pointe d'une distorsion aléatoire. Il comprend également un dispositif de comptages d'erreurs et une gamme de vitesse qui permet le fonctionnement jusqu'aux vitesses télégraphiques normales. La gamme de vitesse de réception maxima est de 5 000 chiffres binaires par seconde, ce qui correspond à de nombreuses vitesses prévisibles de transmission de données.

La précision exceptionnelle de l'instrument en fait un outil de laboratoire fort utile pour les fabricants de matériel de transmission de données et de calculatrices dont les appareils comprennent des dispositifs d'entrée et de sortie pour la transmission de données.

Le contrôleur L.675 est automatiquement protégé contre les connexions erronées à une alimentation secteur d'une tension inappropriée. Lorsque l'instrument est débranché, le réglage de tension secteur revient automatiquement à sa coulée de tension la plus élevée.

Le L.675 et son alimentation stabilisée sont logés dans un robuste coffret métallique ne mesurant que 40,64 cm × 27,94 cm × 22,22 cm. Il est muni d'une courroie de transport, et le panneau frontal ainsi que les boutons de commande, les instruments de mesure et les indicateurs sont protégés par un couvercle frontal à charnières qui s'intercale dans le coffret lorsque l'instrument n'est pas utilisé.

**EE 89 765** *pour plus amples renseignements*

### MOTEURS DE PRÉCISION À TENSION CONTINUE

**Distributeurs: Leland Leroux Ltd, 145 Grosvenor Road, Westminster, London, S.W.1**

*(Illustration à la page 51)*

Ces moteurs de précision miniatures à tension continue, fabriqués par la société Brion Leroux & Cie. de Paris et conçus essentiellement pour la mesure et l'intégration de précision, pour les mécanismes d'asservissement, pour la télécommande et comme générateurs tachymétriques, sont vendus maintenant dans le Royaume-Uni par la société Leland Leroux Ltd.

Ces moteurs sont réalisés en deux dimensions mécaniques de base, et dans chaque cas ils sont livrables en plusieurs types différents. Leur vitesse est en proportion précise de la tension appliquée à leur bobinage; lorsqu'ils sont utilisés comme générateurs, ils produisent une tension proportionnelle au taux de tours. Ils sont d'une efficacité très élevée et se distinguent, en outre, par leur faible courant de démarrage, leur inertie et leur constante de temps très faible. Il existe une gamme de démultiplicateurs pour chaque modèle.

Le plus petit modèle, le Microtor, est livrable en deux types: le type I, de 3 000 tours/min à 5 V et d'une intensité au démarrage de 100 μV; le type II, de 3 000 tours/min à 10 V et d'une intensité au démarrage de 160 μV. Des boîtes de vitesse avec des rapports de démultiplication de 1:18, 1:50, et 1:150 sont fournies pour ce modèle.

Le plus grand modèle, le Birotax, comporte deux bobinages indépendants, pouvant être utilisés en série ou séparément (l'un comme moteur et l'autre comme générateur). Pour une vitesse de 3 000 tours/min le type I exige 50 V sur chaque bobinage; le type VI demande 20 V sur chaque bobinage; le type V, qui est d'une résistance particulièrement basse et qui a été prévu pour les circuits à transistors, exige 5 V sur chaque bobinage; le type II est d'un modèle spécial ayant deux bobinages différents: l'un pour 100 volts comme moteur et l'autre de 9,5 V pour servir de générateur tachymétrique. Des boîtes de vitesse avec des taux de démultiplication de 1:18, 1:50, 1:150, 1:2 000 et 1:3 000 sont livrables pour ce modèle.

**EE 89 766** *pour plus amples renseignements*

### PROGRAMMATEUR VARIABLE

**Brensal Electronics Ltd, Charles Street, Bristol, 7**

*(Illustration à la page 52)*

Le programmateur à contrôle variable Mark III, est un instrument entièrement constitué de corps solides et représente une nouvelle conception d'instruments de contrôle.

Le programmateur Mark III a, en effet, été conçu pour le contrôle numérique de machines-outils ou pour le contrôle de processus où se produit fréquemment la nécessité d'effectuer des changements rapides dans un programme de commutation complexe. Un des nombreux avantages de cet instrument est que n'importe quel changement peut être effectué dans le programme sans interrompre en aucune manière le processus soumis au contrôle; de plus le programmateur peut commuter des canaux à des vitesses beaucoup plus élevées que celles du contrôleur à came de type classique ou de la minuterie électromécanique.

Un programme complet de tous les canaux peut être établi sur plaquette en quelques secondes par des opérateurs non-qualifiés. La souplesse d'emploi de cet instrument est telle que le programme peut être établi de manière à durer un temps total de quelques secondes ou, en ralentissant la fréquence de l'impulsion principale, il peut être porté à plusieurs heures ou même à plusieurs jours.

Le programmateur Mark III que l'on voit dans la photographie, comprend 24 canaux de commutation modulaire, n'importe lequel de ces canaux ou la totalité pouvant être normalement ouvert ou normalement fermé. Ces canaux peuvent être commutés à n'importe quel moment ou plusieurs fois pendant le programme complet.

Le programmateur Mark III a 9 999 incréments auxquels on peut ajuster n'importe lequel ou l'ensemble des 24 canaux pendant le programme. Pour obtenir une résolution plus précise il faut ajouter d'autres décades de module au compteur, pour qu'il puisse fournir n'importe quel nombre voulu d'impulsions pendant n'importe quelle durée de temps. Le nombre de canaux de commutation peut également être augmenté par éléments modulaires jusqu'à au moins 144 pour répondre aux besoins de n'importe quelle application.

**EE 89 767** *pour plus amples renseignements*

## CONTRÔLEUR DE VIBRATIONS

**Dawe Instruments Ltd, Western Avenue, Acton, London, W.3**

*(Illustration à la page 52)*

Le contrôleur de vibrations Dawe type 1435 est un dispositif assez simple conçu pour contrôler les vibrations de manière continue dans presque tous les domaines de la technique moderne. Il a été spécialement conçu à l'origine pour un contrat de centrale nucléaire mais une version commerciale a maintenant été mise au point. Le contrôleur se compose d'un amplificateur électronique à circuit imprimé et à transistors. Il fournit une sortie électrique continue proportionnelle au déplacement de pointe de la structure contrôlée, dérivant d'un transducteur sensible à la vitesse, tel que le Dawe type 1403/1, monté sur un point déterminé de la structure. Le modèle standard, tel que fourni suivant le contrat de la centrale nucléaire, comprend deux canaux séparés, enfermés avec l'alimentation de courant et les relais de surcharge dans un seul coffret compact pour montage mural. C'est là un arrangement pratique permettant aux canaux de se contrôler mutuellement, mais il existe également les éléments à un seul canal ou à plusieurs canaux dont les coffrets et les supports sont conformes aux demandes des clients.

Chaque canal fournit un signal de sortie pour l'indication d'un instrument de mesure à distance. Cette indication est proportionnelle au déplacement de pointe dans la gamme de fréquence de 15 Hz à 800 Hz et elle est plus que suffisante pour la plupart des vibrations mécaniques. On peut également introduire des filtres lorsqu'une gamme de fréquences plus réduite est exigée. La gamme de déplacement pouvant être contrôlée s'étend de 0,001 pouce à 0,010 pouce de pointe suivant le transducteur utilisé, le signal de sortie du moniteur étant de 1 mA au maximum. L'impédance de sortie est de 1 kΩ.

Les relais de surcharge peuvent être préréglés pour fonctionner à n'importe quel niveau de 25% à 100% de la sortie maxima. Dans le modèle standard, ces relais ont une valeur nominale de 10 A 230 V c.a. et ils peuvent être normalement fermés ou normalement ouverts. Ils conviennent donc pour la mise en oeuvre de presque n'importe quel type de circuit avertisseur de surcharge. En séparant le relais du circuit amplificateur, on permet à l'appareil de continuer le contrôle même après qu'un signal d'alarme ait été donné.

Le signal de sortie du contrôleur peut être mesuré par un instrument de mesure à un point de contrôle central éloigné. En variante, il existe un instrument de contrôle local. Dans des conditions d'emploi normal, le contrôleur ne devrait pas exiger de soins autres que la vérification d'étalonnage annuelle une fois que le relais de surcharge a été réglé.

**EE 89 768** *pour plus amples renseignements*

## MACHINE D'ÉTAMAGE

**Fry's Metal Foundries Ltd, Tandem Works, Merton Abbey, London, S.W.19**

*(Illustration à la page 52)*

La machine Flowsolder Roltinner a été conçue pour fournir à l'utilisateur de circuits imprimés une méthode peu onéreuse de protection des plaquettes de circuit. Les circuits de cuivre poli sont revêtus d'un alliage de Flowsolder qui peut être emmagasiné pendant une longue période de temps et assure un meilleur soudage après l'introduction des composants.

Le Roltinner se caractérise par le fait qu'il peut recevoir des plaquettes de circuits imprimés de n'importe quelle longueur et d'une largeur allant jusqu'à 33 cm et d'une épaisseur pouvant atteindre 0,95 cm.

Le principe de base de la machine consiste à faire passer la plaquette de circuit à travers deux rouleaux: un rouleau supérieur et un rouleau inférieur. Le rouleau supérieur est revêtu d'un blindage en acier inoxydable et peut être réglé pour maintenir l'espacement voulu entre le rouleau et la plaquette. Ce rouleau est à ressort de manière à assurer une pression constante. Le rouleau inférieur tourne dans un bain d'alliage Flowsolder.

A mesure que la plaquette de circuit passe entre les deux rouleaux, elle est automatiquement revêtue d'alliage Flowsolder à la vitesse de 182 cm par minute.

La machine est actionnée par un moteur à vitesse constante et à engrenage réducteur. La puissance d'entrée est de 7A à 240 V 50 Hz. Le poids opérationnel est de 126 kg.

**EE 89 769** *pour plus amples renseignements*

## CONDENSATEURS ÉLECTROLYTIQUES MINIATURE

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

*(Illustration à la page 53)*

La société Mullard Ltd vient de créer une nouvelle série de condensateurs électrolytiques miniature de longue durée pour les applications industrielles. Coûtant moins cher que les condensateurs de la série C427, la nouvelle série, type C428, offre de plus grands services tout en conservant les mêmes caractéristiques de durée et de fiabilité que la gamme C427.

Sa durée de vie type à 40° C est de 200 000 heures de fonctionnement continu pour un taux estimé de panne de 0.025% seulement, pour 1 000 heures à pleine tension nominale.

L'emploi de matériaux d'une grande pureté dans la fabrication de ces condensateurs leur assure un faible courant de fuite. Les connexions soudées réduisent la résistance en série au minimum.

La gamme de capacité de la série C428 s'étend aux mêmes valeurs que la série C427, à savoir 2,5 μF à 160 μF cette dernière valeur pouvant être étendue à 320 μF. Les tensions de régime vont de 4 V à 64 V. Les condensateurs se conforment au numéro de groupe climatique IEC 40/70/56.

**EE 89 770** *pour plus amples renseignements*

## CONDENSATEURS AU MICA ARGENTÉ

**London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6**

*(Illustration à la page 53)*

Pour répondre à la demande accrue de condensateurs au mica argenté, une nouvelle gamme de condensateurs moulés vient d'être réalisée. Le premier des éléments de cette gamme est le type 611SM, qui est maintenant en pleine production et livrable à bref délai.

Ce condensateur est basé sur la même construction à lame argentée frittée que le modèle 1106S, couvrant la gamme de 10 pF à 1500 pF à 350 V et de 10 pF à 500 pF à 750 V. Moulé en résine d'époxyde, il répond à la spécification d'humidité DEF5011, catégorie H5, et ne mesure que 15,4 mm × 10,5 mm × 4 mm, les câbles étant placés à entr'axes de 10 mm pour montage sur plaquette de circuit imprimé.

Un processus perfectionné de revêtement d'argent, breveté dans le monde entier, est utilisé sur les lames au mica argenté LEMCO, qui donne une plus longue durée de vie et une plus grande stabilité cyclique. Ce nouveau procédé a été contrôlé selon les normes les plus sévères et il améliore les caractéristiques déjà excellentes du condensateur fritté de base 1106S.

**EE 89 771** *pour plus amples renseignements*

## CONVERTISSEUR ANALOGIQUE/NUMÉRIQUE

**Joyce, Loebl & Co. Ltd, Electron House, Princesway, Team Valley, Gateshead, 11, Co. Durham**

*(Illustration à la page 53)*

Cet appareil constitue une méthode simple pour l'enregistrement automatique de grandes quantités de données finies, aux fins de traitement ultérieur. N'importe quelle suite d'événements représentés par des quantités analogues, par un mouvement mécanique, par la résistance, par le voltage, etc, peut être mesurée, enregistrée en série et dans l'ordre des événements et présentée numériquement. Il peut être ajouté facilement à une gamme étendue d'instruments scientifiques, tels que spectrophotomètres, pH-mètres, etc pour donner une indication numérique, dans la plupart des cas sans modification de l'instrument original.

C'est un instrument à bas prix, d'un encombrement réduit et constitué de circuits à corps solides et à films minces d'un fonctionnement sûr. Le transducteur d'entrée fournit une impédance ne dépassant pas 5 kΩ. Lorsque l'entrée de l'instrument est une tension continue, l'impédance de sortie de la source ne doit pas dépasser 5 kΩ.

Les tensions d'entrée sont converties en fréquencee dans la gamme de 100 à 1 000 Hz, avec une précision de conversion supérieure à 0,1 per cent. (On peut compenser les non-linéarités à des fréquences inférieures à 100 Hz.) Les tensions sont intégrées pendant des périodes allant jusqu'à une seconde et les fréquences sont comptées à travers un compteur de décades à trois étages. Les signaux d'entrée de −12 V mettent en action des compteurs à deux étages pour l'enregistrement de parcours et d'événements dans la série. Des décades supplémentaires peuvent être ajoutées pour enregistrer jusqu'à un maximum de 999 séries et 999 événements dans n'importe quelle série.

Le réglage de zéro est ajusté à partir du panneau frontal de l'instrument et la dérive est de ±1 cycle par heure, c'est à dire supérieure à 1 chiffre après la période initiale de chauffe. L'instrument comprend un potentiomètre incorporé pour la démultiplication dans la gamme de 0 à 999. Le réglage de gamme du convertisseur peut être disposé pour assurer l'enregistrement sur la totalité de l'échelle pour des excursions dans la gamme de zéro à 999.

L'imprimeur de sortie est un élément standard commercial, le ADDO type No. X41E, avec dix positions de tête imprimeuse et une durée d'impression de ligne de moins de deux secondes permettant une période d'intégration d'une seconde par mesure. Lorsque les valeurs mesurées dépassent le cadre du facteur de gamme et d'échelle, une astérisque est imprimée dans la colonne No. 1, les positions de tête imprimeuse restantes étant distribuées comme suit: trois chiffres pour la lecture en série, trois chiffres pour les événements et trois chiffres pour la valeur mesurée. Une version du convertisseur est en cours de mise au point et donnera soit des sorties d'affichage soit des sorties sur bande perforée.

**EE 89 772** *pour plus amples renseignements*

### RELAIS EN CASCADE

**Flight Refuelling Ltd, Wimborne, Dorset**

*(Illustration à la page 53)*

Le relais en cascade FR, type RSC-51, mis au point par la Division d'électronique industrielle de la société Flight Refuelling Ltd, présente des qualités de compacité supérieures à celles de son prédécesseur, le RSC-41. Dans les montages de relais en cascade utilisant un certain nombre de modules RSC-51, chaque impulsion d'entrée fait avancer le relais en cascade d'une position et sa vitesse de fonctionnement est beaucoup plus élevée que celle d'un sélecteur mécanique d'une durée de vie sans entretien de 100 millions d'opérations.

Chacune des positions est pourvue de deux contacts de sortie entièrement isolés et d'un commutateur à lame vibrante Hamlin, ce qui assure aux modules tous les avantages du commutateur à lame vibrante.

Le RSC-51 peut être actionné à partir d'un transistor à faible puissance, si nécessaire, et il ne produit aucune interférence de radiation. Il mesure 1,30 pouce de côté × 0,83 pouce de profondeur.

Le courant constant nominal est de 10,9 mA, la vitesse de fonctionnement maxima est de 150 impulsions par seconde et la durée minimum de l'impulsion d'entraînement est de 2 msec.

**EE 89 773** *pour plus amples renseignements*

# Résumés des Principaux Articles

**Un générateur de signaux aléatoires** par D. A. G. Tait et M. Skinner

*Résumé de l'article aux pages 2 à 7*

*Les sources de bruit conventionnelles produisent une sortie pouvant être considérée comme étant due à une infinité de phénomènes infinitésimaux (par exemple, les collisions d'ions dans un tube à gaz). Le nouvel instrument sont il est question dans cet article permet de synthétiser le signal à partir d'un nombre contrôlé et déterminé de sources de paramètres statistiques bien définis. Il en résulte un signal d'une souplesse remarquable et parfaitement stationnaire.*

**Un amplificateur de puissance transistorisé à large bande** par R. C. French

*Résumé de l'article aux pages 8 à 11*

*Cet article décrit un amplificateur de puissance transistorisé à large bande dont le gain est de 27dB, la largeur de bande de 3dB allant de 2MHz à 60MHz et l'excursion de sortie de 22 volts à une pointe de 75Ω à 30MHz. La puissance de sortie est de 0,8 watts.*

**Un module pour éléments microminiature** par A. J. Mayhew

*Résumé de l'article aux pages 12 à 20*

*Après avoir analysé les méthodes courantes de construction d'instruments numériques, l'auteur décrit les caractéristiques exigées d'un module idéal et présente une nouvelle réalisation qui répond à ces caractéristiques. Il montre qu'une densité d'emmagasinage élevée peut être obtenue au moyen d'un câblage distinct et de composants circulaires et que des modules compatibles peuvent utiliser des circuits à plusieurs couches. Il décrit différentes variétés, anciennes et nouvelles, de ces circuits ainsi que l'application de l'automatisation à la conception, à la construction et au contrôle de ces circuits. Les possibilités de développement du nouveau module sont considérables.*

**Un oscillateur à tension contrôlée et à circuit intégré standard** par A. D. Walker

*Résumé de l'article aux pages 21 à 23*

*L'auteur décrit un circuit simple pouvant produire un rapport entre l'entrée et la sortie, rapport qui demeure linéaire en fonction de la période ou de la fréquence. La base du circuit est un amplificateur différentiel à circuit intégré standard. Ce circuit assure un fonctionnement satisfaisant jusqu'à 100kHz. Il est d'un encombrement et d'un poids réduits, sa consommation électrique est faible et il est d'une très grande fiabilité.*

**Un amplificateur à courant continu et à large bande pour charge inductive** par M. Bronzite

*Résumé de l'article aux pages 24 à 30*

*Il s'agit ici d'un amplificateur c.c. à réponse linéaire allant jusqu'à 100kHz et pouvant être utilisé avec une charge inductive. L'amplificateur de tension comporte une réaction locale qui stabilise le gain du circuit ouvert. Il comporte également une réaction de mode commun pour assurer la stabilité thermique ainsi qu'une réaction générale pour linéariser la réponse. L'amplificateur de courant a aussi un circuit pour empêcher les fuites thermiques. L'auteur de l'article analyse le problème de la stabilité de l'amplificateur aux fréquences élevées et étudie en détail une méthode permettant de résoudre ce problème. Il examine, en outre, la stabilité thermique ainsi que les conditions d'alimentation en tension continue qui ont pu être réalisées d'une manière nouvelle.*

**Modulation de code à impulsions binaires et à pente quantifiée** par M. N. Faruqui

*Résumé de l'article aux pages 31 à 35*

*Une nouvelle méthode a été réalisée pour le codage des pentes de signaux à l'aide d'impulsions de 1/0 à un seul chiffre, ainsi que d'un circuit à réaction spéciale et d'un seuil dans le comparateur de code. On a constaté que le comportement général du système est meilleur sous plusieurs rapports que celui des systèmes de modulation Delta et Delta-Sigma. La gamme dynamique des systèmes a été étendue par l'emploi d'un compresseur-extenseur instantané mais le système n'utilise que des circuits simples. Les caractéristiques du système sont analysées et comparées, et il est démontré que ce système, à 40KHz de fréquence de répétition des impulsions, est équivalant sous tous les rapports au modèle familier de modulateur de code d'impulsions à 6 chiffres.*

**Un amplificateur à transistors opérationnel sans correction de dérive automatique** par M. Carapic et D. Velasevic

*Résumé de l'article aux pages 36 à 38*

*Cet article traite de la conception d'un amplificateur à transistor opérationnel sans correction automatique de glissement. Ce dernier est, en effet, très faible par suite de l'emploi de paires équilibrées de transistors bipolaires dans deux étages différentiels. La dérive de l'amplificateur, utilisé comme inverseur, est de 150μV pour une gamme de température de 30°C et la largeur de bande est de 600kHz.*

**Une méthode simplifiée de réalisation des circuits multivibrateurs à transistors** par P. Mars

*Résumé de l'article aux pages 38 à 39*

*Cet article décrit une méthode de réalisation de circuits multivibrateurs à transistors ayant des valeurs spécifiques de durée et d'amplitude d'impulsion, de durée de rétablissement de l'état initial et de stabilité de température du temps de détente. Etant donné qu'on ne tient pas compte des composants réactifs de l'impédance d'entrée du transistor, le processus de réalisation ne s'applique au circuit multivibrateur qu'avec des temps de détente relativement étendus. Les transistors conducteurs sont censés être saturés.*

**Le bruit du transistor semi-conducteur à oxyde de métal** par S. M. Bozic

*Résumé de l'article aux pages 40 à 41*

*Les résultats expérimentaux de l'indice de bruit du transistor semi-conducteur à oxyde de métal sont indiqués dans la gamme de fréquence de 30 à 1600kHz et pour une résistance de source de 600Ω à 100kΩ. L'effet du point de fonctionnement sur l'indice de bruit est également indiqué. Les conclusions établies sont brièvement discutées.*

**Un oscillateur à cristal utilisant des amplificateurs à circuit intégré** par A. H. James

*Résumé de l'article aux pages 42 à 43*

*Il s'agit dans cet article d'un oscillateur utilisant un cristal de quartz comme élément de résonance série en liaison avec des amplificateurs constitués de corps solides. Il en résulte un circuit simple et compact dont la stabilité de fréquence est de 60 parties par million entre −50°C et +100°C et de 0,5 parties par million par volt de changement de haute tension.*

# NEUE AUSRÜSTUNGEN

*Übersetzung der Seiten 48 bis 53*

**Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.**

### NF Mikrovoltmeter

**Dymar Electronics Ltd, Rembrandt House, Whippendell Road, Watford, Hertfordshire**

*(Abbildung Seite 48)*

Das NF-Mikrovoltmeter 705, das neuste Instrument im Dymar-Einschubsystem, ist ein empfindliches Röhrenvoltmeter zum Messen von Wechselspannungen im Tonfrequenzbereich, dessen Frequenzgang jedoch mit voller Genauigkeit bis zu 300 kHz geht.

Ein Merkmal des Gerätes ist der symmetrische Eingang, der direktes Messen von Gegentaktspannungen gegen Erde gestattet.

Die Gleichtaktunterdrückung ist über den grössten Teil des Frequenzbereiches höher als 60 dB, was das Messen erdsymmetrischer Spannungen in Gegenwart viel grösserer Störspannungen gestattet. In bezug auf Erde unsymmetrische Spannungen können an jeden der Eingänge gelegt werden.

Die Eingangsempfindlichkeit ist 100 µV für Vollausschlag mit einem unter 20 µV liegenden Eigenrauschpegel; dieser Rauschpegel lässt sich durch Beschränkung der Bandbreite auf 10 kHz auf 8 µV reduzieren.

Die Eingangsimpedanz ist 10 MΩ für jeden Eingang, und vorhandene Ausgangsklemmen gestatten Verwendung des Gerätes als empfindlicher Vorverstärker für Oszillografen oder Endverstärker.

**EE 89 751** *für weitere Einzelheiten*

### Breitbandverstärker

**Telequipment Ltd, 313 Chase Road, Southgate, London, N.14**

*(Abbildung Seite 48)*

In ihrem Programm für Verstärkereinschübe stellt Telequipment Ltd nunmehr den Typ H vor, der Breitbandanforderungen entspricht und besonders für das Messen von Anstiegszeiten geeignet ist.

Zwei durch Schalter wählbare Eingangsbuchsen sind 60 dB getrennt. Die Bandbreiten sind 0...25 MHz (−3 dB) bei Ablenkfaktoren von 100 mV/cm... 50 V/cm und 0...5 MHz (−3 dB) bei Ablenkfaktor 10 mV/cm. Der Ablenkfaktor wird in jedem Fall im Verhältnis 1:2:5 geschaltet, und ein Verstärkungsregler gibt kontinuierliche Regelung zwischen den Abschwächerstufen.

Der Verstärker H ist genau so wie die anderen fünf von der Firma hergestellten Verstärkereinschübe zwischen den Telequipment-Oszillografen Typen S43, D43 und D43R sowie dem gerade eingeführten Typ D53 austauschbar.

**EE 89 752** *für weitere Einzelheiten*

### Tester für Oberflächenbeschaffenheit

**Payne Products International Ltd, Buckingham Avenue, Trading Estate, Slough, Buckinghamshire**

*(Abbildung Seite 48)*

Die Hersteller der Lapmaster-Flachläppmaschinen, Payne Products International Ltd, kündigen einen neuen Mikrotester—den Diavite Baumuster II—für das Messen aller anerkannten Bereiche der Oberflächenbeschaffenheit an. Das mit einphasigem Wechselstrom von 110...220 V betriebene neue Modell kann Mittelliniendurchschnitte zwischen 0 und 0,0076 mm messen. Die Skala ist durch Drucktasten in vier Bereiche unterteilt: bis zu 254 µm, 762 µm, 0,0025 mm und 0,0076 mm.

Das Gerät hat zwei Grenzwerte, und zwar 0,076 mm für sehr feine konkave oder konvexe Profilabschnitte, oder 0,76 mm für flache Profilabschnitte.

Ein neues Konstruktionsmerkmal ist eine Warnlampe für Toleranzen, die auf Grenzwerte des Mittelliniendurchschnittes oder $R_{max}$ eingestellt werden können, für jeden zu testenden Teil zugelassen sind und bei deren Überschreiten ein rotes Licht aufleuchtet.

Neu ist ausserdem, dass für den Hub des Messstiftes drei verschiedene Längen vorgesehen sind: 10 mm, 20 mm und 4 mm. Die Messstiftgeschwindigkeit ist 1,27 mm/s. Das neue Modell ist tragbar und wiegt nur 6 kg.

**EE 89 753** *für weitere Einzelheiten*

### Dekadenzählermoduln

**Quarndon Electronics Ltd, Slack Lane, Derby**

*(Abbildung Seite 48)*

Diese Baureihe von Dekadenzählermoduln ist für industrielle Systeme und Laboranwendung gedacht, für die der Aufwand für komplette elektronische Zähler nicht gerechtfertigt ist. Diese modularen Einheiten stellen ein äusserst anpassungsfähiges System für Spezialaufgaben dar.

Die Einheiten haben Frontplattenabmessungen von nur 121 × 38 mm und sind 140 mm tief. Anschluss erfolgt über eine 32polige Standardsteckverbindung für gedruckte Schaltungen, und die Zählanzeige erscheint auf einer Standard-Zählröhre.

Moduln können entweder mit Silizium- oder Germanium-Halbleitern bestückt geliefert werden. Die maximale Zählgeschwindigkeit des DCM501 ist 1 MHz. Andere Germanium-Bausteine sind für Zählgeschwindigkeiten von 5 MHz und 12 MHz lieferbar. Die Siliziummoduln haben Zählgeschwindigkeiten von 200 kHz, 2 MHz und 30 MHz. Es gibt auch umkehrbare Zähler. Für Darstellung des Kommas gibt es eine extern betätigte Miniaturneonlampe.

Die Art des erforderlichen Eingangsimpulses hängt in gewissem Masse von seiner Anstiegszeit ab; für Höchstgeschwindigkeit muss der Impuls eine Spannung von 2,5 V und eine Anstiegszeit von 500 ns haben.

Der Ausgang des Moduls kann einen weiteren Modul treiben, so dass mehrstellige Zähler aufgebaut werden können.

Für Fernanzeige usw. gibt es 8-4-2-1-binärkodierte Dezimalausgänge.

Durch Anlegen eines Löschimpulses an entsprechende Klemmen kann man die Einheit auf Null zurückstellen. Das Programm wird durch Kristall-Frequenznormale, Ausblendschaltungen und Fernanzeigen vervollständigt, so dass komplette Zähler zusammengestellt werden können.

**EE 89 754** *für weitere Einzelheiten*

### Robuste Reflexklystrons

**The English Electric Valve Co. Ltd. Chelmsford, Essex**

(*Abbildung Seite 49*)

In Verfolg der Forderungen nach relativ preiswerten, doch robusten Überlagerern mit breiter Auswahl von Frequenztemperaturbeiwerten hat English Electric Valve Co Ltd die Fertigung einer neuen Baureihe kompakter X-Band-Reflexklystrons aufgenommen.

In der Herstellung dieser neusten Röhrengruppe finden dieselben strengen Produktionsmethoden Anwendung, die für den Bau der robusten EEV-Klystrons für militärische Zwecke vorgeschrieben sind. Ausnutzung der dabei gewonnenen Erfahrungen ergibt als sofort sichtbare Vorteile höhere Leistungsfähigkeit und Zuverlässigkeit der neuen Röhren.

In dieser Klystronserie hat man die Möglichkeit, für jeden bestimmten Anwendungszweck eine Röhre mit geeignetem Frequenztemperaturbeiwert zu wählen. Typ K300 z.B. hat einen Beiwert von −130 kHz/°C, der dem Beiwert der meisten kommerziell greifbaren X-Band-Klystrons entspricht. Ausserdem sind diese neuen Röhren in der Fertigung weitgehend gegeneinander austauschbar, so dass die für die einzelnen Anwendungen geeignetsten Parameter für Klystrons in Sonderausführung einfacher permutiert werden können.

Ein typisches Beispiel ist Typ K3003, der elektrisch mit der Ausführung K3020 mit freien Anschlüssen identisch ist. Typ K3003 ist ideal für Marinezwecke geeignet und besonders durch seine ausgezeichneten Frequenzdrifteigenschaften gekennzeichnet. In Langzeittests von über 3000 Stunden blieben die Änderungen innerhalb 10 MHz. Ausserdem ist die im Pflichtenblatt festgelegte Höchstfrequenzmodulation unter Schwingungsbelastung von 10 g maximal 200 kHz.

**EE 89 755** *für weitere Einzelheiten*

### Mikrologik-Schulung

**Lan-Electronics Ltd, 97 Farnham Road, Slough, Buckinghamshire**

(*Abbildung Seite 49*)

Das Mikrologik-Schulungsgerät LD-20 ist eine Entwicklung des bekannten LD-15 mit integrierten Halbleiterschaltungen für die Logik. Bei Anwendung dieser Moduln besteht die Aufgabe des Elektronikers nicht mehr im Einzelentwurf der Schaltungen, sondern in den Verbindungen zwischen Schaltungen. Das Mikrologik-Schulungsgerät gibt dem zukünftigen Anwender integrierter Halbleiterschaltungen die Möglichkeit, nach allen Gesichtspunkten dieser Technik praktische Erfahrungen zu sammeln. Jede integrierte Halbleiterschaltung besteht aus einer kompletten digitalen oder linearen Schaltung, in der alle Schaltungselemente aus einem kleinen Halbleiter-Einzelscheibchen hergestellt und in ein Transistorgehäuse (wie z.B. das TO-5) oder ein Flachpaket (mit typischen Abmessungen von 6,6 × 6,6 mm und 1,3 mm dick) eingekapselt werden.

Die integrierten Halbleiterschaltungen des LD-20 werden mit einer Speisespannung von nur 3,6 V betrieben und haben einen Stromverbrauch von unter 1 mA, was für den Trend in Logik nach niedrigen Pegeln typisch ist. Diese niedrigen Leistungspegel sind augenfällige Vorteile bei der Entwicklung kompakter Ausrüstungen.

Ausser den fünfzehn integrierten Halbleitergattern gibt es im LD-20 fünf Trennverstärker zum Treiben der Anzeigelampen. Der LD-20 hat ausser der Fernsprechwählerscheibe des LD-15 für Eingabezwecke fünf zusätzliche Kippumschalter, wodurch ihm grössere Vielseitigkeit gegeben wird. Diese Schalter legen an nebenliegende Buchsen wahlweise positive oder negative Spannungen für die Steckverbindung zur Logik. Trotzdem die logischen Schaltungen grundsätzlich als NOR-Gatter entworfen sind (mit 0 V = 0 und Speisespannung = 1) können diese fünfzehn Gatter bei Vorzeichenumkehr durch Umlegen der Eingangskippschalter für die NAND-Funktion eingesetzt werden.

Das Mikrologik-Schulungsgerät hat eine eingebaute Stromversorgung mit Netzanschluss, die die korrekte Gleichspannung abgibt, wird mit dem LANDEC-Handbuch für digitale Schulungsrechner (über 100 Seiten) und dem von S.G.S.-Fairchild bearbeiteten Mikrologik-Handbuch sowie einem kurzen, speziell für das Mikrologik-Schulungsgerät LD-20 geschriebenen Handbuch geliefert.

**EE 89 756** *für weitere Einzelheiten*

### Höchstspannungs-Konstantstromversorgung

**Cathodeon Electronic Ltd, Bircham Road, Southend-on-Sea, Essex**

(*Abbildung Seite 49*)

Eins der ersten Geräte, das der neugegründete Geschäftsbereich Präzisionsmessungen der Cathodeon Electronic Ltd herausgebracht hat, ist die 5-kV-Konstantstromversorgung.

Das Gerät gibt entweder positive oder negative Ausgangsspannung in bezug auf Masse bei jeder vorgewählten Stromentnahme bis zum vollen Nennwert von 5 mA ab. Die Vorwahl erfolgt auf der Frontplatte.

Ein transistorisierter Spezialmodul gibt vollen Überlastungsschutz und kann auf Ansprechen im Strombereich 2...5 mA eingestellt werden. Der Kurzschlussstrom ist nur 20 Prozent höher als die Einstellung für die Überlastungsschwelle.

Die Regelung der Stromversorgung ist über den vollen Spannungsbereich besser als 0,1 Prozent, und das mit Doppelskala ausgerüstete Messgerät ist auf Strom oder Spannung umschaltbar.

Anwendungsgebiete für das Modell PMD 101 bestehen als Stromversorgung für Photovervielfacher, als Höchstspannungsquelle für Elektronenstrahlröhren usw.

Das PMD 101 wird in einem Gehäuse für Arbeitsplatzeinsatz oder für Einbau in ein 19″-Gestell geliefert.

**EE 89 757** *für weitere Einzelheiten*

### Temperaturregler

**C.N.S. Instruments Ltd, 61 Holmes Road, London, N.W.5**

(*Abbildung Seite 50*)

C.N.S. Instruments Ltd hat einen neuen Temperaturregler herausgebracht. Das Gerät trägt die Typenbezeichnung S.R.2. und ist ein volltransistorisierter proportionaler Temperaturregler hoher Genauigkeit mit magnetischen Verstärkern.

Zu den verschiedenen neuartigen Merkmalen gehört eine einem Leistungsverhältnis von 30:1 entsprechende Strombegrenzungseinrichtung, die die Regelung von Platin- und Molybdänöfen ermöglicht.

Mittels eines Umschalters lässt sich entweder der vom Regler abgegebene Leistungsanteil oder die Diagonalspannung der Platinthermometer-Messbrücke anzeigen und dadurch die Abweichung der Temperatur vom Sollwert schätzen. Eingebaute Kontrollen erlauben wiederholtes Testen bei einer Temperatur mit äusserst geringer Unsicherheit.

Die angewendete Druckplattentechnik gestattet schnelles Auswechseln des grössten Teils der Schaltungen bei Ausfall eines Bauelementes.

Das Instrument ist für Anschluss rein ohmscher oder induktiver Verbraucher geeignet; die die tatsächliche Ofenleistung regelnden robusten magnetischen Verstärker können übermässige Spannungsstösse und Stromüberlastung aushalten.

Der S.R.2 ist mit dem 50gängigen C.N.S. - Präzisions - Schleifdrahtpotentiometer ausgerüstet, das mit $2 \times 10^{-4}$ Unsicherheit wiedereinstellbar ist, so dass die Einstellung der Solltemperatur äusserst genau erfolgen kann.

An einer typischen kleineren Ofeninstallation unter normalen Bedingungen und bei nichtstabilisierter Stromversorgung ist die Regelung bei 1000°C und mit einem C.N.S. Platinthermometer innerhalb ±0.02 Prozent konstant. Automatische Netzspannungskompensation, die innerhalb weniger Perioden auf Netzschwankungen anspricht, ist eingebaut.

Die niedrige Energieaufnahme (40 VA) gestattet dichtes Stapeln ohne übermässiges Überhitzen, so dass keine Druckkühlung erforderlich ist.

**EE 89 758** *für weitere Einzelheiten*

### Digitalvoltmeter

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

*(Abbildung Seite 50)*

Das neue Dynamco-Messgerät DM2006 ist ein Festkörper-Digitalvoltmeter, dessen neue Spannungsumwandlungstechnik in bezug auf Schaltungselemente äusserst wirtschaftlich ist. Dadurch kann Dynamco (früher Digital Measurements Ltd) ein preisgünstiges Voltmeter produzieren, ohne hohe Leistungsfähigkeit und Konstruktionsqualität einzubüssen.

Das Gerät besteht aus einem Mess- und Anzeigegrundgerät mit Skala bis zu 9999, dessen Bereich durch Einsteckmoduln bestimmt wird. Der einfachste Einschub hat vier Bereiche von 100 $\mu$V... 1 kV mit 0,05 Prozent Anzeigeunsicherheit. Ausgangssignale zum Treiben von Ausgabegeräten können ohne Benachteiligung der Gleichtaktunterdrückung des Gerätes durch Einsteckkarten vorgesehen werden. Moduln mit höherer Empfindlichkeit, Bereichsautomatik, sowie zum Messen von Wechselspannungen und Widerständen sind geplant und werden mit Weiterentwicklung des Programmes lieferbar werden.

Zuverlässigkeit wird besonders betont, und für den Bau des DM2006 werden z.B. gedruckte Schaltungen auf glasfaserverstärkten Epoxydplatten, Siliziumtransistoren und durchweg für militärische Anwendung zugelassene Bauelemente verwendet.

**EE 89 759** *für weitere Einzelheiten*

### Zwillingstiefpass

**The Wayne Kerr Laboratories Ltd. Sycamore Grove, New Malden, Surrey**

*(Abbildung Seite 50)*

Ein neues NF-Filter von Wayne Kerr besteht aus zwei identischen Kanälen, die sich entweder einzeln, in Kaskade oder in Bandpassanordnung verwenden lassen. Ein einziger Abstimmknopf überstreicht 28 Bereiche und gestattet genaues Einstellen beider Kanäle auf jede gewünschte Frequenz zwischen 0,1592 Hz und 1592 Hz.

Als Zwillingstiefpass ist das Modell SA500 besonders für Leistungsmessungen an Steuerungs- und Regelsystemen geeignet. Ein Filter wird in die Ausgangsleitung des Systems geschaltet, um den Rauschabstand zu verbessern. Das zweite Filter wird in den Bezugssignalpfad geschaltet, um identische Kennlinien einzuführen und dadurch die Gesamtübertragungsfunktion des Systems vom Filter unabhängig zu machen.

Bei Kaskadenschaltung der beiden Tiefpässe ist die Dämpfung 36 dB/Oktave. In der Bandpassschaltung ist die Verstärkung 1:1, die Phasenverschiebung bei der Mittenfrequenz Null und die Dämpfung auf jeder Seite 18 db/Oktave.

Der Eingangswiderstand jedes Kanals ist 2 M$\Omega$ (ohmisch) und der Ausgangswiderstand unter 60 $\Omega$. Das Modell SA500 ist für Gestelleinbau oder in einem tragbaren Gehäuse lieferbar.

**EE 89 760** *für weitere Einzelheiten*

### Verzerrungsmessgerät

**Standard Telephones & Cables Ltd. S.T.C. House, 190 Strand, London, W.C.2**

*(Abbildung Seite 50)*

Das Verzerrungsmessgerät 74834-A wurde für dynamische Wobbelmessungen an Mikrowellen- und anderen Fernmeldesystemen eingeführt und misst die durch Nichtlinearität des Systems hervorgerufenen Amplituden- und Phasenverzerrungen des übertragenen Signals. Zusammen mit einem geeigneten Oszillografen kann das Instrument differentielle Amplituden von 0,05...100 Prozent und differentielle Phasen von 0,02°...90° messen.

Vier alternative HF-Testsignale stehen zur Verfügung; eins davon ist die Fernseh-Farbhilfsträgerfrequenz von 4,4297 MHz, die drei anderen sind 304 kHz, 2 MHz und 7 MHz. Andere Frequenzen können zur Befriedigung von Sonderanforderungen vorgesehen werden. Jedes der HF-Testsignale kann mit dem niederfrequenten Wobbelsignal mit vorwählbarer Amplitude kombiniert werden, so dass das Signalgemisch nacheinander die verschiedenen Betriebsamplitudenbereiche des Systems durchläuft. Die Wobbelfrequenz ist 81 Hz. Marken auf dem Bildschirm des Oszillografen dienen der vereinfachten Ablesung der Amplituden- und Phasenverzerrung.

Das Gerät kann für Schleifenmessungen oder—mit einem anderen gleichen Gerät—für Streckenmessungen eingesetzt werden; andererseits kann man den Empfänger an einem Ende der Strecke verwenden und den aus dem Gehäuse ausgebauten Generator am anderen zusammen mit einer Stromversorgung und einem Monitorgerät, die als Zubehör lieferbar sind.

**EE 89 761** *für weitere Einzelheiten*

### Fotoelektrischer Reflexionsbaustein

**Lancashire Dynamo Electronic Products, Rugeley, Staffordshire**

*(Abbildung Seite 50)*

Das Fertigungsprogramm der LDEP (MI-Gruppe) für fotoelektrische Ausrüstungen wurde mit der Einführung eines neuen Tastkopfes Baureihe RPH6 für Einsatz mit ihren transistorisierten Steuergeräten erweitert.

Der nach dem Reflexionsprinzip arbeitende RPH6 besteht aus einem fotoelektrischen Ferntastkopf, in dem sowohl die Strahler- wie auch die Empfängerelemente untergebracht sind. Die Elemente sind so angeordnet, dass die Empfängerzelle den vom Zielobjekt reflektierten Lichtstrahl des Strahlers "sieht".

Die Ausrüstung ist für alle Anwendungsmöglichkeiten geeignet, bei denen es auf Positionieren, Zählen, Schutz gegen Überschiessen und ähnliche Anforderungen ankommt.

Der in Leichtmetallspritzguss ausgeführte Tastkopf besteht aus zwei Hauptteilen: dem Gehäuse und dem Deckel. Der abnehmbare Deckel gestattet Zutritt zum Inneren ohne Störung der optischen Elemente. Eine Gummidichtung macht den Kopf wetterfest, und die Konstruktion ermöglicht starren Einbau, der Fluchtungsfehler durch Stoss oder Schwingungen ausschliesst.

Der Reflexionszusammenbau besteht aus einer Standard-8-W-Lichtquelle, einem fokussierenden Glasreflektor von 38 mm Durchmesser × 38 mm und einer Zellenhalterung.

Der Reflektor sitzt in einer Aluminiumhalterung mit Vorrichtungen für horizontale und vertikale Regelung sowie Verschiebung nach vorn und hinten, um gute Schärfe des Heizfadenbildes über den ganzen Bereich von etwa 50 cm bis zur grössten Betriebsstrecke zu gewährleisten.

Für Abtasten über kürzere Entfernungen, oder wenn ein schärferes Heizfadenbild erforderlich ist, kann eine zusätzliche Linse von 38 mm Durchmesser eingebaut werden.

Durch Zusammenbau verschiedener optischer und elektrischer Elemente in verschiedene Kombinationen kann der Typ RPH6 den unterschiedlichen Einsatzbedingungen angepasst und über Entfernungen von etwa 2,5 cm...61 m betrieben werden.

Der Grundbaustein ist mit einer Halbleiterfotozelle ORP.60 bestückt, die ihren Widerstand ändert, wenn sie Licht ausgesetzt wird, und bei Umgebungstemperaturen bis zu 45°C zufriedenstellend arbeitet.

**EE 89 762** *für weitere Einzelheiten*

### Impulsgeber

**K.S.M. Electronics Ltd. 139-149 Fonthill Road, Finsbury Park, London, N.4**

*(Abbildung Seite 51)*

Der Impulsgeber T-15 ist ein äusserst kompaktes, mit Qualitäts-Bauelementen gebautes Festkörpergerät in Druckschaltungstechnik. Trotz seiner Kompaktheit ist das Instrument sehr vielseitig und gibt Einzel- und Doppelimpulse mit gleichzeitiger positiver und negativer Polarität ab. Impulsbreiten und -verzögerungen von 30 ns...30 ms lassen sich einstellen, und der interne Oszillator hat einen Frequenzbereich von 15 Hz ...

15 MHz. Impulsbreiten bis zu fast 100% Tastverhältnis können erreicht werden. Die Amplitudenregelung ist kontinuierlich, und positive oder negative Impulse mit Amplituden bis zu 10 $V_s$ werden erzeugt.

Das Gerät kann durch den eignen internen Oszillator, durch interne "Einmal"-Drucktaste oder externe Wellenform getriggert werden. Stromtor- oder Wobbelbetrieb von externen Signalen ist ebenfalls möglich.

**EE 89 763** *für weitere Einzelheiten*

### Relaisverzögerungsbaustein

**Vertrieb: D. Robinson & Co. Ltd, Victoria House, 44 Park Street, Camberley, Surrey**

*(Abbildung Seite 51)*

Clifford & Snell Ltd hat zwei neuentwickelte elektronische Verzögerungsbausteine für das Einsteck-Relaissystem D2600 bekanntgegeben.

Der erste ist eine Ausführung des 'Superslug A' für Betrieb mit 110 V~. Die Abbildung zeigt, dass dieser Baustein äusserlich einem Relais D2600 ähnlich ist und in eine Fassung gleicher Art eingesteckt wird.

Er wird Superslug genannt, weil er das Abfallen eines zugeordneten Relais "slugs" oder verzögert, und zwar auf ähnliche Weise wie ein Kupferring, aber auf viel längere Zeit.

Eine aus einer Spannungsverdopplerschaltung gespeiste Kaltkathodenröhre gibt Verzögerungen von 1...72 s mit einer Wiederholgenauigkeit von etwa ±1 Prozent.

Auf Zuverlässigkeit wurde in der Konstruktion besonderer Wert gelegt, und der Baustein wird überall dort Anwendung finden, wo Clifford & Snell-Relais eingesetzt werden, d.h. in Schiffen, Kraftwerken usw. Zu den Schutzmassnahmen gehören eine interne Sicherung und ein Begrenzungswiderstand.

Der zweite neue Baustein ist der "Superslug-B2", der durch Änderung der Anschlüsse entweder mit 24 V— oder 50 V— betätigt wird. Er ersetzt den Superslug-B nach Konstruktionsverbesserungen und kostet etwa ein Drittel weniger.

**EE 89 764** *für weitere Einzelheiten*

### Daten-Testgerät

**The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex**

*(Abbildung Seite 51)*

Dieses neue Gerät Typ L.675 wurde vom Geschäftsbereich Fernmeldetechnik der M.E.L. entwickelt, um den Bedarf an einem tragbaren Gerät zur Leistungsmessung von Datenübertragungsausrüstungen und -kreisen zu befriedigen. Es entspricht dem Pflichtenblatt der britischen Postbehörde für den Datel-Tester 1A.

Das Instrument hat getrennte Sende- und Empfangsteile, so dass die Sendung über einen Kanal einer Wechselverkehrsanlage gesendet und über den anderen Kanal mit unterschiedlicher Geschwindigkeit empfangen werden kann.

Merkmale des Daten-Testgerätes sind grosses, eindeutiges und direktanzeigendes Messgerät zum Messen der mittleren Neutralverzerrung und Spitzenwerte der Zufallsverzerrungen, eine Fehlerzähleinrichtung und ein Geschwindigkeitsbereich, der Betrieb bis zu normalen Telegrafiegeschwindigkeiten herunter erlaubt. Die höchste Geschwindigkeit auf der Empfangsseite ist 5000 Bits/Sek, so dass der Bereich viele der voraussehbaren Datenübertragungsgeschwindigkeiten umfasst.

Durch die ungewöhnliche Genauigkeit eignet sich das Gerät auch für Laborarbeiten bei Herstellern von Datenübertragungsausrüstungen und Rechneranlagen, in deren Systemen Ein- Ausgabegeräte für Datenübertragung vorgesehen sind.

Das Gerät ist automatisch gegen unbeabsichtigten Anschluss an die falsche Netzspannung geschützt. Nach Abschalten geht der Spannungswähler automatisch auf den Abgriff für die höchste Spannung zurück.

Typ L.675 und seine Konstantstromversorgung sind in einem robusten Metallgehäuse mit Abmessungen von nur 406 × 280 × 222 mm untergebracht, das mit Tragriemen geliefert wird. Frontplatte, Bedienknöpfe und Anzeiger werden durch einen Klappdeckel geschützt, den man während des Einsatzes im Gehäuse unterbringen kann.

**EE 89 765** *für weitere Einzelheiten*

### Präzisions-Gleichstrommotoren

**Vertrieb: Leland Leroux Ltd, 145 Grosvenor Road, Westminster, London, S.W.1**

*(Abbildung Seite 51)*

Diese von Brion Leroux & Cie in Paris hergestellten Präzisions-Gleichstromkleinmotoren wurden im wesentlichen für Präzisionsmessungen und Integration, für Servoeinrichtungen, Fernsteuerung sowie als tachometrische Generatoren entwickelt und können nunmehr in England durch Leland Leroux Ltd bezogen werden.

Diese Motoren werden in zwei Grundgrössen hergestellt; für beide gibt es verschiedene unterschiedliche Typen. Ihre Geschwindigkeit ist genau der an ihre Wicklungen angelegten Spannung proportional; bei Einsatz als Generatoren ist die abgegebene Spannung der Umlaufgeschwindigkeit proportional. Sie haben einen sehr hohen Wirkungsgrad, niedrigen Anlaufstrom, sehr niedrige Trägheit und Zeitkonstante. Für jedes Modell gibt es eine Auswahl von Untersetzergetrieben.

Das kleinere Modell—der Microtor—ist in zwei Typen lieferbar: Typ I gibt bei 5 V 3000 UPM und benötigt ein Anlaufmoment von 100 $\mu$W; Typ II gibt 3000 UPM bei 10 V und erfordert ein Anlaufmoment von 160 $\mu$W. Für dieses Modell sind Getriebe mit Untersetzungsverhältnissen von 1:18, 1:50 und 1:150 lieferbar.

Das grössere Modell—der Birotax—hat zwei getrennte Wicklungen, die in Serie oder getrennt (eine als Motor, eine als Generator) zu benutzen sind. 50 V sind an jeder Wicklung von Typ I erforderlich, um 3000 UPM zu erzeugen; Typ VI benötigt 20 V an jeder Wicklung, während Typ V einen besonders niedrigen Widerstand für Transistorschaltungen hat und 5 V an jeder Wicklung benötigt. Typ II ist ein Spezialtyp mit zwei verschiedenen Wicklungen: eine für 100 V als Motor, die andere für 9,5 V für Einsatz als tachometrischer Generator. Getriebe mit Untersetzerverhältnissen von 1:18, 1:50, 1:2000 und 1:3000 sind für dieses Modell lieferbar.

**EE 89 766** *für weitere Einzelheiten*

### Veränderliches Programmiergerät

**Brensal Electronics Ltd. Charles Street, Bristol, 7**

*(Abbildung Seite 52)*

Dieses Gerät wird als Programmiergerät für veränderliche Steuerung Baumuster III bezeichnet, ist durchweg in Festkörpertechnik ausgeführt und stellt ein neues Konzept in der Konstruktion von Steuergeräten dieser Art dar.

Das Programmiergerät wurde für Anwendung in der numerischen Werkzeugmaschinensteuerung und Verfahrenstechnik entwickelt, bei denen die Ausführung schneller Änderungen in komplexen Schaltprogrammen häufig erforderlich ist. Einer der vielen Vorteile des Instrumentes ist die Möglichkeit beliebige Programmänderungen vorzunehmen, ohne dabei in irgendwelcher Weise den gesteuerten Prozess aufzuhalten. Ausserdem können Kanäle mit viel höherer Geschwindigkeit umgeschaltet werden als bei herkömmlichen Nockensteuerungen oder elektromechanischen Schaltuhren.

Auf einem Steckfeld oder mittels Knopfschalter kann ungelerntes Personal in nur wenigen Sekunden ein komplettes Programm für alle Kanäle einstellen. Das Gerät ist so anpassungsfähig, dass eingestellte Programme eine Gesamtablaufzeit von nur wenigen Sekunden haben können oder—durch Verlangsamung der Taktfrequenz—sich auf mehrere Stunden oder sogar mehrere Tage ausdehnen lassen.

Das abgebildete Programmiergerät Baumuster III hat 24 modulare Schaltkanäle; jeder beliebige oder alle Kanäle können als Arbeits- oder Ruhekontakte ausgeführt und während des Gesamtprogrammes zu beliebiger Zeit oder mehrmals betätigt werden.

Baumuster III hat 9999 Inkremente,

auf die jeder beliebige oder alle Kanäle während des Programmes eingestellt werden können. An den Zähler lassen sich weitere dekadische Moduln anbauen, um ein höheres Auflösungsvermögen zu erreichen, so dass für jede vorgegebene Zeit jede erforderliche Impulsanzahl möglich ist. Die Anzahl der modularen Schaltkanäle kann auch bis auf mindestens 144 erhöht werden, um den Anforderungen fast jeder Anwendungsmöglichkeit zu genügen.

**EE 89 767** *für weitere Einzelheiten*

### Schwingungsmonitor

**Dawe Instruments Ltd, Western Avenue, Acton, London, W.3**

*(Abbildung Seite 52)*

Der neue Dawe-Schwingungsmonitor 1435 ist eine einfache Einrichtung zur laufenden Überwachung von Schwingungen in fast allen technischen Sektoren. Das ursprünglich für einen Kernkraftwerkauftrag entwickelte Instrument liegt jetzt in kommerzieller Ausführung vor. Es besteht aus einem elektronischen Verstärker, der transistorisiert und in Druckschaltungstechnik gebaut ist und von einem geschwindigkeitsempfindlichen Abtaster (Messwandler)—z.B. dem Dawe 1403/1—gespeist wird. Das kontinuierliche Ausgangssignal des Verstärkers ist der Spitzenverschiebung der überwachten Struktur proportional. Die für das Kernkraftwerk gelieferte Standardausführung hat zwei getrennte Kanäle, die zusammen mit Stromversorgung und Überlastungsrelais in einem kompakten Gehäuse für Wandmontage untergebracht sind. Das ist eine bequeme Ausführung, in der die beiden Kanäle sich gegenseitig kontrollieren; Ein- und Mehrkanalgeräte sind jedoch auch lieferbar und können in Gehäusen und für Montage nach Kundenwünschen gefertigt werden.

Jeder Kanal gibt ein Ausgangssignal für Fernanzeige ab, das der Spitzenverschiebung im Frequenzbereich 15 ... 800 Hz, das den grössten Teil der mechanischen Schwingungen erfasst, proportional ist. Für enger begrenzte Frequenzbereiche lassen sich auf Wunsch Filter einschalten. Der Verschiebungsmessumfang hängt vom Messwandler ab; im allgemeinen können Spitzenwerte von 0,025 bis zu 0,25 mm überwacht werden. Das maximale Ausgangssignal vom Monitor ist 1 mA, der Ausgangswiderstand 1 kΩ.

Die Überwachungsrelais können für Auslösung bei jedem Pegel zwischen 25 Prozent und 100 Prozent des Höchstausgangsstroms eingestellt werden. In der Standardausführung sind sie für 10 A 230 V~ bemessen und sowohl als Arbeits- wie auch Ruhekontakte lieferbar. Sie sind daher für Alarmschaltungen oder Überlastungsabschaltung fast jeder Art geeignet. Die Trennung des Relais von der Verstärkerschaltung ermöglicht Fortsetzung der Überwachung nach Auslösung des Überlastungsrelais.

Das Ausgangssignal kann mittels eines Messgerätes in einer zentralen Warte gemessen werden. Andererseits sind auch Messgeräte für Messung an Ort und Stelle lieferbar. Unter angemessenen Bedingungen erfordert der Monitor nach Einstellung des Überlastungsrelais ausser der jährlichen Prüfung der Eichung keinerlei Wartung.

**EE 89 768** *für weitere Einzelheiten*

### Verzinnungsmaschine

**Fry's Metal Foundries Ltd, Tandem Works, Merton Abbey, London, S.W.19**

*(Abbildung Seite 52)*

Die Flowsolder-Roltinner-Machine bietet den Anwendern gedruckter Schaltungen ein billiges Verfahren zum Schutz der Leiterplatten. Die gereinigten Kupferleiter werden mit Flowsolder-Legierung überzogen, wodurch eine längere Lagerfähigkeit und besseres Löten nach Bestücken mit Bauelementen erreicht wird.

Ein Hauptkonstruktionsmerkmal des Roltinners ist die Fähigkeit, Leiterplatten bis zu 33 cm Breite, fast jeder Länge und bis zu 9,5 mm Stärke zu bearbeiten.

Dem Arbeitsprinzip entsprechend laufen die Leiterplatten zwischen zwei Walzen—einer oberen und einer unteren —durch die Maschine. Die obere Walze ist mit rostfreiem Stahl ummantelt und kann so eingestellt werden, dass zwischen Walze und Plattenstärke der richtige Abstand besteht. Die obere Walze ist gefedert, so dass sie einen konstanten Druck ausübt. Die untere Walze läuft in einem Flowsolder-Bad.

Eine durch beide Walzen laufende Leiterplatte wird automatisch mit einer Geschwindigkeit von etwa 180 cm/min mit Flowsolder-Legierung überzogen.

Die Maschine wird durch einen untersetzten, gangkonstanten Motor getrieben. Die Energieaufnahme ist 7 A bei 240 V, 50 Hz und das Betriebsgewicht 127 kg.

**EE 89 769** *für weitere Einzelheiten*

### Elektrolyt-Kleinkondensatoren

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

*(Abbildung Seite 53)*

Eine neue Baureihe von Elektrolyt-Kleinkondensatoren mit langer Lebensdauer für industrielle Anwendungszwecke wurde von Mullard angekündigt. Die neue Baureihe C428 ist billiger als die C427-Serie, bietet ein grösseres CV-Produkt in derselben Gehäusegrösse und hat gleichzeitig in bezug auf Betriebslebensdauer und Zuverlässigkeit dieselben Eigenschaften wie die bestehende Baureihe C427.

Typische Lebensdauerwerte sind 200 000 Stunden bei Dauerbetrieb und 40°C mit einer geschätzten Ausfallrate von nur 0,025 %/1000 h bei voller Nennspannung.

Verwendung hochreiner Werkstoffe in der Fertigung der Kondensatoren sichert einen niedrigen Reststrom, und geschweisste Anschlüsse halten den Serienwiderstand niedrig.

Baureihe C428 ist in denselben Kapazitätswerten lieferbar wie Baureihe C427, d.h. 2,5 μF bis zu 160 μF mit einer Erweiterung bis auf 320 μF. Betriebsspannungen sind von 4 V bis 64 V. Die Kondensatoren genügen den IEC-Vorschriften für Klimaklasse 40/70/56.

**EE 89 770** *für weitere Einzelheiten*

### Silber-Glimmerkondensatoren

**London Electrical Manufacturing Co. Ltd, Bridges Place, Parsons Green Lane, London, S.W.6**

*(Abbildung Seite 53)*

Zur Befriedigung des steigenden Bedarfes an Silber-Glimmerkondensatoren wurde eine völlig neue Baureihe mit Kunststoff umpresster Kondensatoren entwickelt. Der erste Typ in dieser neuen Baureihe ist der jetzt in voller Produktion und kurzfristig lieferbare 611SM.

Der Kondensator beruht auf demselben Konstruktionsprinzip wie der 1106S, d.h. dem gesinterten, versilberten Blatt, und ist in Werten von 10 pF bis zu 1500 pF bei 350 V und 10 pF bis zu 500 pF bei 750 V lieferbar. Die Schutzhülle besteht aus Epoxydharz und entspricht der Feuchtigkeitsklasse H5 nach DEF 5011. Die Abmessungen sind 15,4 mm × 10,5 und 4,0 mm mit Anschlussdrähten in 10 mm Abstand für Bestückung von Druckschaltungskarten.

Ein durch Weltpatente geschütztes neues Versilberungsverfahren wird jetzt für LEMCO-Silber-Glimmerblätter angewendet und ergibt bessere Lebensdauer und zyklische Stabilität. Das neue Verfahren wurde unter strengen Bedingungen erprobt und hat bewiesen, dass es die bereits ausgezeichneten Eigenschaften der gesinterten Kondensatoren 1106S in ihrer Grundkonstruktion steigert.

**EE 89 771** *für weitere Einzelheiten*

### Analog-Digitalwandler

**Joyce, Loebl & Co. Ltd, Electron House, Princesway, Team Valley, Gateshead, 11. Co. Durham**

*(Abbildung Seite 53)*

Die automatische Aufzeichnung grosser Mengen endlicher Information für nachfolgende Verarbeitung ist mit diesem Gerät sehr einfach. Jede beliebige, durch analoge Mengen, mechanische Bewegung, Widerstand, Spannung usw. dargestellte Vorgangsfolge kann gemessen, serienweise und in Vorgangsfolge erfasst und digital dargestellt werden. Das Gerät lässt sich ohne Schwierigkeiten an eine breite Auswahl bestehender wissenschaftlicher Instrumente, Spektrophotometer, pH-Messer usw. anschliessen, um mit ihnen—in den meisten Fällen ohne Änderung des Originalgerätes—digital drucken oder darstellen zu können.

Das preiswerte Gerät hat kleine Abmessungen und ist durch Bestückung mit Festkörper- und Dünnfilmschaltungen im Betrieb zuverlässig. Die Impedanz der Eingangswandler ist nicht höher als 5 kΩ. Wo das Eingangssignal in das Gerät aus einer Gleichspannung besteht, soll die Ausgangsimpedanz der Quelle 5 kΩ nicht überschreiten.

Eingangsspannungen werden in Frequenzen im Bereich 100...1 100 Hz umgesetzt, wobei die Umsetzunsicherheit besser als 0,1 Prozent ist. (Nichtlinearitäten werden in die Frequenzen unter 100 Hz versetzt.) Spannungen werden in Perioden bis zu einer Sekunde integriert und Frequenzen mittels eines dreistufigen Dekadenzählers erfasst. Eingangssignale von $-12$ V lösen zweistufige Zähler zur Erfassung von Durchlaufreihen und Vorgängen innerhalb dieser Reihen aus. Extradekaden zum Registrieren von Reihen bis zu einem Maximum von 999 und bis zu 999 Vorgängen in beliebigen Reihen können angebaut werden.

Nulleinstellung erfolgt von der Frontplatte des Gerätes aus, und die Drift ist nach der anfänglichen Anheizperiode $\pm 1$ Periode je Stunde, d.h. besser als eine Ziffer. Für Untersetzen des Instrumentes innerhalb des Bereiches 0...999 ist ein eingebautes Potentiometer vorhanden. Die Bereicheinstellung des Wandlers kann so ausgelegt werden, dass sie bei Aufzeichnung mittels Schreiber und Auslenkungen innerhalb des Bereiches 0...999 die volle Skala durchläuft.

Der ADDO-Typ X41E, ein kommerziell greifbares Gerät, ist ein Ausgangsdruckwerk mit zehn Druckpositionen und einer Zeilendruckzeit von weniger als zwei Sekunden, wodurch eine Integrationsperiode von einer Sekunde je Messung möglich wird. Wenn die Messwerte ausserhalb des Bereiches und Untersetzungsfaktors fallen, wird in Kolonne 1 ein Sternchen gedruckt. Die anderen Kolonnen sind wie folgt zugeteilt: drei Stellen für die Reihennummer, drei für den Vorgang und drei für den Messwert. In einer in Vorbereitung befindlichen Ausführung des Wandlers kann der Ausgang entweder angezeigt oder als Lochstreifen abgegeben werden.

**EE 89 772** *für weitere Einzelheiten*

**Schrittschalter**

**Flight Refuelling Ltd, Wimborne, Dorset**

(*Abbildung Seite 53*)

Der Geschäftsbereich Industrielle Elektronik der Flight Refuelling Ltd bietet jetzt den FR-Schrittschalter RSC-51 an, der gegenüber seinem Vorgänger RSC-41 eine viel bessere Packdichte aufweist. In Schrittschalteranordnungen mit einer Anzahl RSC-51-Moduln verursacht jeder Eingangsimpuls das Vorrücken des Schrittschalters um eine Stellung, und zwar mit einer Geschwindigkeit, die ein Vielfaches mechanischer Schrittschalter ist, und einer wartungsfreien Lebensdauer von bis zu 100 Millionen Schaltvorgängen.

Jede Stellung ist mit zwei völlig isolierten Ausgangskontakten bestückt, und zwar handelt es sich jeweils um einen Hamlin-Zungenschalter, so dass jeder Modul alle Vorteile eines Schutzgas-Zungenschalters hat.

Ein Transistor geringer Leistung kann den RSC-51 ohne Erzeugung von Störstrahlungen treiben. Die Abmessungen sind 33 × 33 × 21 mm tief.

Der Nennhaltestrom ist 10,9 mA, die höchste Betriebsgeschwindigkeit 150 i.p.s. und die Mindestbreite des Treibimpulses 2 ms.

**EE 89 773** *für weitere Einzelheiten*

# Zusammenfassung der wichtigsten Beiträge

**Ein Rauschgenerator** von M. Skinner und D. A. G. Tait

*Zusammenfassung des Beitrages auf Seite 2-7*

*Herkömmliche Rauschquellen haben ein Ausgangssignal, das man als auf unendlich viele, unendlich kleine Vorgänge (z.B. den Zusammenstoss von Ionen in einer Gasröhre) zurückführbar betrachten kann. In einem neuen, hier beschriebenen Gerät wird das Signal von einer endlichen und geregelten Anzahl von Quellen gut definierter statistischer Parameter zusammengesetzt, was ein Signal ungewöhnlicher Vielseitigkeit und feststehender Verteilungskurve ergibt.*

**Ein transistorisierter Breitband-Leistungsverstärker** von R. C. French

*Zusammenfassung des Beitrages auf Seite 8-11*

*Ein beschriebener transistorisierter Breitband-Leistungsverstärker hat eine Verstärkung von 27 dB, eine 3-dB-Bandbreite von 2 MHz . . . 60 MHz und eine Ausgangsspannung von 22 Vss an 75 Ω bei 30 MHz. Die Ausgangsleistung ist 0,8 W.*

**Ein Baustein für Mikrominiaturelemente** von A. J. Mayhew

*Zusammenfassung des Beitrages auf Seite 12-20*

*Nach kritischer Betrachtung der in der Digitaltechnik vorherrschenden Technologie werden die an einen idealen Modul zu stellenden Anforderungen gegeben und ein neuer Baustein, der ihnen entspricht, beschrieben. Es wird gezeigt, dass mit diskreter Verdrahtung und kreisrunden Bauelementen hohe Packdichten erreicht werden können und kompatible Moduln mit Mehrebenenschaltungen möglich sind. Verschiedene alte und neue Abarten der letzteren, ebenso wie der Einfluss der Automatisierung auf ihren Entwurf, die Konstruktion und Tests werden kurz besprochen. Es wird gezeigt, dass dieser Baustein bedeutende Entwicklungsmöglichkeiten hat.*

**Ein spannungsgesteuerter Oszillator mit integrierter Standardschaltung** von A. D. Walker

*Zusammenfassung des Beitrages auf Seite 21-23*

*Eine beschriebene einfache Schaltung ruft zwischen Eingang und Ausgang eine Beziehung hervor, die entweder mit Periode oder Frequenz linear ist. Die Schaltung beruht auf der integrierten Standardschaltung eines Differentialverstärkers. Bei kleinen Abmessungen, niedrigem Gewicht, geringer Energieaufnahme und hoher Zuverlässigkeit arbeitet die Schaltung zufriedenstellend bis zu 100 kHz.*

**Ein Breitband-Gleichstromverstärker für induktive Belastung** von M. Bronzite

*Zusammenfassung des Beitrages auf Seite 24-30*

*Der Beitrag beschreibt einen für induktive Belastung geeigneten Gleichstromverstärker mit linearem Frequenzgang bis zu 100 kHz. Der Spannungsverstärker hat örtliche Gegenkopplung zur Stabilisierung der Leerlaufverstärkung, Rückkopplung für thermische Stabilität und Gesamtgegenkopplung zur Linearisierung des Frequenzganges. Der Stromverstärker hat auch eine Schaltung zur Verhinderung thermischen Durchgehens. Die Verstärkerstabilität bei hohen Frequenzen wird untersucht und eine Methode zur Minderung dieses Problems eingehender besprochen. Die thermische Stabilität wird weiter besprochen und ausserdem Anforderungen an die Gleichstromversorgung, denen in etwas neuartiger Weise genügt wird.*

**Eine flankenquantisierte, binäre Pulscodemodulation** von M. N. Faruqui

*Zusammenfassung des Beitrages auf Seite 31-35*

*Nach einem neuentwickelten Verfahren werden Signalflanken mit einstelligen 1/0 Impulsen kodiert, wobei eine Spezialrückkopplungsschaltung und ein Schwellenwert im Übersetzer-Komparator Verwendung finden. Die Gesamtleistung des Systems ist in vieler Hinsicht besser als die der Delta- und Delta-Sigma-Modulationssysteme. Die Dynamik des Systems wurde durch Verwendung eines momentanen Kompandors erweitert, jedoch finden im System nur einfache Schaltungen Anwendung. Die Eigenschaften des Systems werden besprochen und verglichen und gezeigt, dass dieses System bei einer Impulsfolgefrequenz von 40 kHz in jeder Beziehung der bekannten 6stelligen PCM gleichwertig ist.*

**Ein Transistor-Funktionsverstärker ohne automatische Driftkorrektur** von M. Carapic und D. Velasevic

*Zusammenfassung des Beitrages auf Seite 36-38*

*Dieser Beitrag befasst sich mit Entwurf und Analyse eines Transistor-Funktionsverstärkers, dessen Drift nicht automatisch korrigiert wird.*

*Niedrige Drift wird durch Bestückung der zwei Differentialstufen mit zusammenpassenden Pärchen bipolarer Transistoren erreicht. Die Drift des als Inverter verwendeten Verstärkers ist für einen Temperaturbereich von 30°C 150 μV, die Bandbreite 600 kHz.*

**Eine vereinfachte Behandlung des Entwurfes von Transistor-Multivibratorschaltungen** von P. Mars

*Zusammenfassung des Beitrages auf Seite 38-39*

*In dem Beitrag wird eine Methode für den Entwurf von Transistor-Multivibratorschaltungen mit kennzeichnenden Werten für Impulsdauer und -amplitude, Erholungszeit des ursprünglichen Zustandes und temperaturkonstanter Relaxationszeit beschrieben. Da die reaktiven Elemente der Transistoreingangsimpedanz vernachlässigt werden, ist dieses Entwurfsverfahren nur für Multivibratorschaltungen mit relativ langer Relaxationszeit zulässig. Es wird angenommen, dass leitende Transistoren gesättigt sind.*

**Rauschen der Metalloxyd-Halbleiter-Transistoren** von S. M. Bozic

*Zusammenfassung des Beitrages auf Seite 40-41*

*Für den Rauschwert von MOS-Transistoren im Frequenzbereich 30 . . . 1600 kHz und den Quellenwiderstand von 600Ω bis zu 100 kΩ werden Versuchsergebnisse gegeben. Weiterhin wird der Einfluss des Arbeitspunktes auf den Rauschwert gezeigt. Die gezogenen Schlussfolgerungen werden kurz besprochen.*

**Ein sin/cos-Funktionsgeber mit hohem Auflösungsvermögen** von S. Harigovindan

*Zusammenfassung des Beitrages auf Seite 42-43*

*Ein neuartiger Funktionsgeber wird beschrieben, der sin/cos $\theta$ als Gleichspannung mit einer Genauigkeit von besser als 0,1 Prozent des Endwertes erzeugt, wenn $\theta$ als binärkodierte Dezimalzahl mit einer Auflösung von 0,1 Prozent im Bereich 0 . . . $2\pi$ eingegeben wird. Die Schaltung besteht aus Präzisions-Spannungsteilerstufen, die durch Relais geschaltet werden. Zwei Ausführungen werden beschrieben.*