# ELECTRONIC ENGINEERING

VOL. 36 No. 434

APRIL 1964

# Commentary

THE Sub-Group of the C.C.I.R. Study Group XI consisting of some 100 delegates from 20 countries together with representatives of private and industrial organizations met in London last February to discuss the present position of colour television and to attempt to reach agreement on a common colour television system for Europe.

The choice of a common system for the European broadcasting area is an important one for not only would it standardize colour television in Western Europe—including the United Kingdom—but also in those Asian and African countries bordering the Mediterranean, and ultimately extending to the remainder of Europe and Asia.

An earlier meeting of this Sub-Group was held, also in London, in the summer of last year when the BBC, the ITA, the Post Office and the television industry went to considerable lengths to give an impressive display of the colour television systems under consideration, but no decision was reached at the time because it was felt that the newer systems needed more time for development.

As is now well known, the choice lies between the American NTSC, the French SECAM and the German PAL systems and it was expected when the meeting closed last summer that the intervening period would allow sufficient time for further experimental work to be carried out.

Without oversimplifying the problem, the position is that the NTSC system has been in existence for ten years or more, its advantages and disadvantages are well known, and the BBC and the British television industry in particular have acquired considerable knowledge and experience as a result of the work that has been done in modifying this system to our standards and, but for the quite recent introduction of the SECAM and PAL systems, the BBC would have adopted the NTSC system and the rest of Europe would have followed suit, for the simple reason that there was no alternative.

The SECAM and PAL systems were developed with the object of improving on the NTSC system but as is so often the case these improvements have introduced other problems so that the choice lies between the NTSC system with its acknowledged limitations and one of the two new, promising, but not fully developed systems.

At the February meeting the delegates had before them considerable evidence of the further field trials and experiments carried out on the three systems by members of the European Broadcasting Union, including France, Holland, Italy, Switzerland and West Germany, the countries immediately concerned, along with the United Kingdom.

These trials and experiments covered all aspects of a public colour television service—the design of studio equipment, tape recording, propagation, transmitter and receiver design, but it appears that the conference was still unable to reach a unanimous decision.

The British delegates representing the BBC, the ITA, the Post Office and industry were whole-heartedly in favour of the NTSC system and their opinions are best summarized in a statement issued by the Post Office which said that "The view of the United Kingdom experts, as endorsed by the Television Advisory Committee, is that, taking all factors into account, the NTSC system is better than the other two systems. The United Kingdom delegation at the C.C.I.R. meeting, therefore pressed for its adoption".

This view was supported by Holland and, not unnaturally, by the United States but the remaining delegates felt that still more time was needed for further study of the problem and the outcome was that it was felt preferable to postpone the decision until the next meeting of Study Group XI which is to be held in Vienna in the spring of next year.

At the summer meeting last year the BBC did not come out in favour of the NTSC system but merely expressed the view that had it then been adopted "it would be possible to introduce colour television on a limited scale early in 1965" so that this further postponement means that colour television in Britain will be delayed for at least three years from now.

The position in this country at the moment, again referring to the official Post Office statement, is that "The future policy to be adopted in this country will now be reviewed in the light of these discussions; the Television Advisory Committee, on which the broadcasting authorities, the radio industry and the Government Departments are represented, is to meet in March to formulate their advice to the Postmaster General."

And at any moment now the Postmaster General must consider the evidence submitted to him by the Television Advisory Committee and decide whether to "go it alone", which presumably means "go it alone" with the NTSC system—or to await the outcome of next spring's meeting in Vienna which almost certainly means either the SECAM or the PAL system and if one can hazard a guess at this early stage this means the SECAM system.

The technical arguments for and against the separate systems are well known but the purely technical considerations alone are by no means the deciding factor and other points of view, political, economic and so on, are becoming more prominent.

It is essential, if live exchange is to be an important feature of colour television, that there should be a common system throughout Western Europe and for that matter throughout Eastern Europe as well and the argument put forward that as this country will, without doubt, be the first country in Europe to introduce colour we can therefore make our own decision and expect the rest to follow does not seem acceptable.

# A Chart Reader and Statistical Analyser

By J. A. Phillips*

*An equipment is described which reads roll charts and derives the mean reading, the variance and a histogram. The scanning system employs a vidicon television tube, is completely automatic and gives a binary-coded output. A small fixed-program computer is used for the statistical computation.*

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

THE problem of analysing a large number of roll charts frequently arises. Charts are often several feet long and the analysis may require the determination of all or any of the following variables—maximum or minimum values, turning points, rates of change, mean reading and standard deviation. An accurate determination of the last two requires a large number of readings to be taken from the chart. As an example, consider a chart fourteen feet long, representing one week's recording at one inch per hour. If the chart is read by eye and the mean is estimated in each quarter-inch length, 672 readings will be taken. At a rate of one reading every ten seconds this relatively coarse analysis will take nearly two hours. These readings must be sorted into a histogram and a considerable amount of arithmetic is then necessary to derive the mean and standard deviation.

The present equipment is designed to derive the mean and variance (standard deviation squared) automatically. Its speed is such that a chart fourteen feet long is processed in about two and a half minutes.

Fig. 1 shows the layout of the various components. The camera, illumination system and chart drive mechanism form a single assembly which stands together with the control unit on a table supported between two cabinets. The computer and power supplies are housed in the left-hand cabinet and the right-hand one is storage space for the charts.

### Principle of the Reader

The chart is wound through the field of view of a simple television camera. This uses a vidicon tube, has a 'raster' of two lines and a scanning frequency of fifty frames per second. A scale of black and white bars is superimposed on the chart, as shown in Fig. 2, the width of the scale coinciding with the working width of the chart. The first line of the raster scans across the chart adjacent to the scale and the second line scans the scale.

The video waveform from the tube is fed to a video amplifier and then to pulse circuits which measure, during the first scan, the time from the start to the trace on the chart. During the second scan the black bars on the scale are counted from the start for this same length of time, and the number of bars counted is a measure of the chart reading. The output is available as a train of pulses or as a binary-coded number. Since both edges of the bars are counted and there are fifty of them the resolution is 100.

This technique eliminates errors due to non-linearity of the scan, drift of the raster or distortion by the lens etc. Many variations are possible. For example the bars may be spaced logarithmically, or according to some other law. Digital circuits are readily devised to recognize limits, rates of change and turning points. Several different scales can be provided and the second scan switched to cover the required one.

### Resolution

The necessary reading resolution is related to the accuracy of the original recording and in this instance 100 parts is sufficient. However various precautions are necessary to achieve this resolution and if these are

Fig. 1. The complete equipment

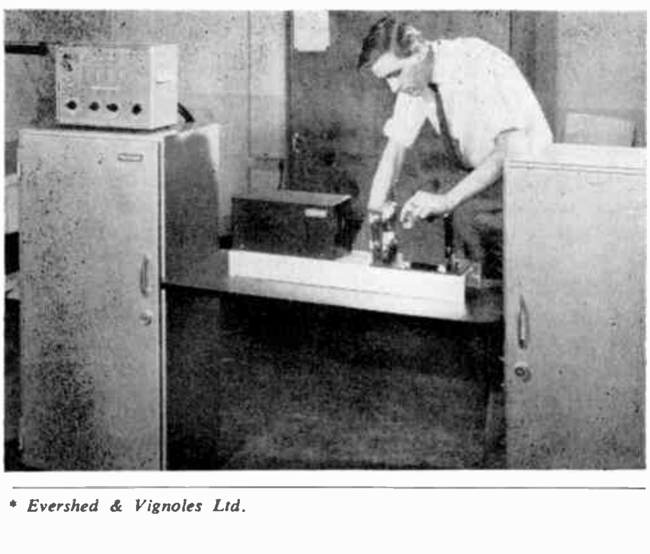
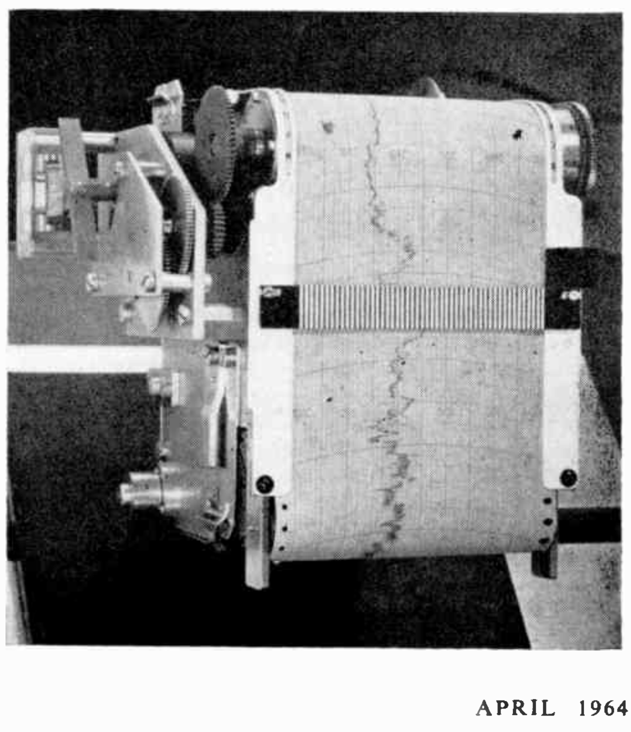


Fig. 2. The chart and scale



* *Evershed & Vignoles Ltd.*

applied to the limit the resolution can be extended to at least 200 parts.

The factor which limits the resolution is the diameter of the spot in the vidicon. This is about one two-hundredth of the available scan length.

The best focus in the tube is obtained when it is operated with a high wall-anode voltage and suitably high focusing and deflexion fields. In practice there is little improvement to be obtained over 500V. The present equipment uses 350V.

Vidicons are now available in which the wall-anode is electrically isolated from the mesh. If there is a small potential difference between these electrodes an improvement of focus is obtained. Such a tube has yet to be tried in this application.

The design of the scanning and focus coils has a great effect on focus. It is important to make the diameter of the coils large, so that the fields are uniform throughout the deflexion area. The speed of scanning is relatively slow and there is no difficulty in providing the necessary scanning power. A minimum diameter of two inches has been found necessary.

in increasing the scanning frequency beyond a certain rate since this merely leads to redundant readings. The maximum chart speed is 1in/sec and the scanning frequency is 50 frames/sec. For reasons which will be discussed in the section on the design of the pulse circuits the actual number of distinct readings obtained is slightly less than this.

The scanning velocity (as opposed to the scanning frequency) has a proportionate effect on the signal from the vidicon and it is desirable to scan as fast as possible in order to reduce the video amplification required. However an upper limit is set by the frequency response of the amplifier and the counting rate of the pulse circuits. In this equipment the scanning velocity is about 9mm/sec, giving a scan time of 1·4msec and a video frequency of 70kc/s when scanning the black bars.

**Chart Illumination**

A high brightness level is desirable so as to obtain a large signal and a short lag with the vidicon. In practice a brightness approximating to daylight can be obtained from tungsten filament lamps without excessive heat dissipation or elaborate focusing arrangements. At this level the vidicon requires a target voltage of 50V.

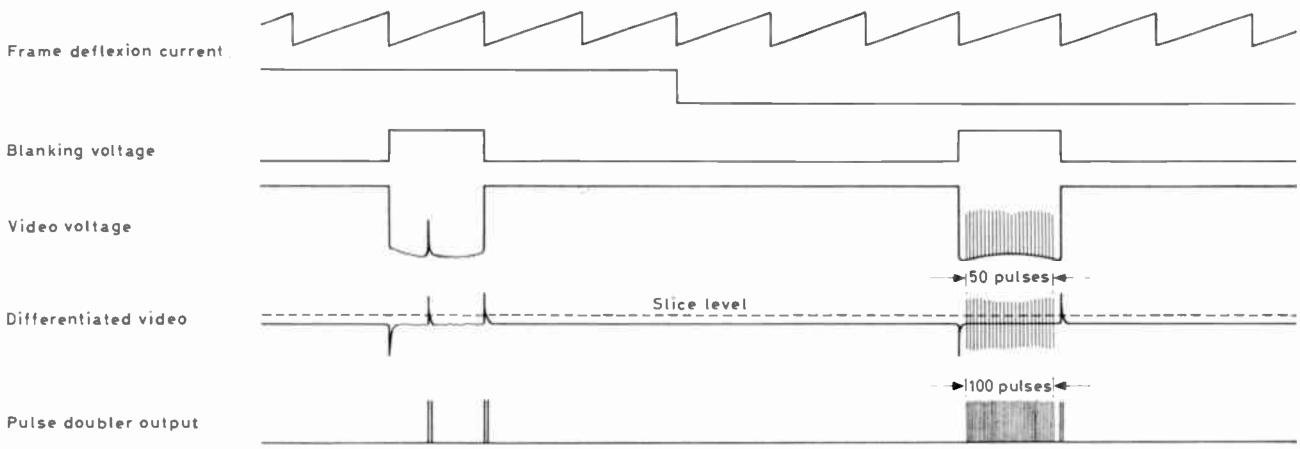



Fig. 3. Camera waveforms

**Design of the Chart**

The charts have a working width of three inches and are made on dry electro-sensitive paper. This paper is light grey in colour but because the trace is dense black there is adequate contrast between the trace and the background.

The signal from the vidicon is seriously reduced when the image of the trace is narrower than the diameter of the scanning spot. This fixes the trace width at not less than 0·015in and there is no difficulty in obtaining this.

A printed scale is required on the chart so that it can be read by eye, and this must be rendered invisible to the camera. The printing is done in light yellow and an orange filter is used on the lens. In predominantly tungsten light complete suppression is obtained, yet the colour is sufficiently dense to be read quite easily by the naked eye in the same light.

**Speed of Reading**

Due to the lag effects in the vidicon there is a limitation to the speed at which an image can be moved across the target of the tube without loss of signal amplitude. This in turn limits the speed at which the chart can be moved if the maximum rates of change of the trace are to be read satisfactorily. The maximum image velocity is about 0·5in/sec.

When the chart speed is limited there is little point

**The Chart Drive Mechanism**

The reading rate is constant at fifty per second and to simplify the computer, the number of readings taken is either 8192 or 1024. Therefore to alter the length of chart read, the speed of the chart is varied. Each speed must be related to the scanning frequency so that the correct length is read. This is achieved by operating the chart drive motor as a phase-lock velodyne with reference frequencies derived from the same oscillator that feeds the scanning circuits.

When short lengths of chart are being analysed the process is speeded up by reducing the number of readings taken from 8 192 to 1 024. The reading times are thus 8 192 ÷ 50 = 162sec and 1 024 ÷ 50 = 20sec.

The drive mechanism comprises an Evershed type FAH a.c. motor-tachometer driving the sprocket rollers through a fixed-reduction gear train. A toothed wheel and reluctance pick-up generate a frequency proportional to the motor speed which is required for the phase-lock circuit. The chart passes vertically past the camera and the drive unit is pivoted so that the chart can be swung into the horizontal plane and away from the lamps to facilitate loading.

The illumination is provided by four 24W bulbs in reflectors.

**The Camera**

The arrangement of the camera tube and the circuits is quite conventional except for the frame time-base and the pulse-forming circuits which follow the video amplifier.

The frame deflexion current waveform is square, and is generated by a switching circuit. The peak amplitudes are separately adjustable, enabling the two lines of the raster to be positioned independently in the vertical direction, so no vertical shift coils are required on the vidicon. Slight overscanning in the horizontal direction avoids the need for horizontal shift coils.

Pulses are produced from the video waveform by differentiating the wave and slicing the resultant signal. The advantage of this method is that the low-frequency components of the waveforms are suppressed. If this were not done it would be necessary to ensure uniform illumination of the chart and to avoid shading and target blemishes in the vidicon. The disadvantage is that the rise-time of signal from the trace is sometimes too long to pass the differentiating circuit. This can be so when the scanning spot crosses the trace obliquely, i.e. when the rate of change of the trace is large. With the particular charts used, a small percentage of readings is lost. This is not serious because of the large number of readings taken and the averaging process to which they are subjected. When a reading is lost the previous reading is repeated to the computer so that the total number of readings remains correct. Fig. 3 shows the video waveforms obtained for each scan line and the corresponding differentiated wave. The slice level is adjusted so that only the pulses corresponding to transitions from white to black are detected. These pulses are passed to a doubling circuit which produces a second pulse shortly after each of the applied pulses. The time interval is arranged to equal the time from one edge of a black bar to the other so that in effect a pulse is produced for each edge of the bar.

The output of the pulse-doubling circuit is passed to the computer cabinet where the reader logic is housed. This logic produces the binary output to feed the computer.

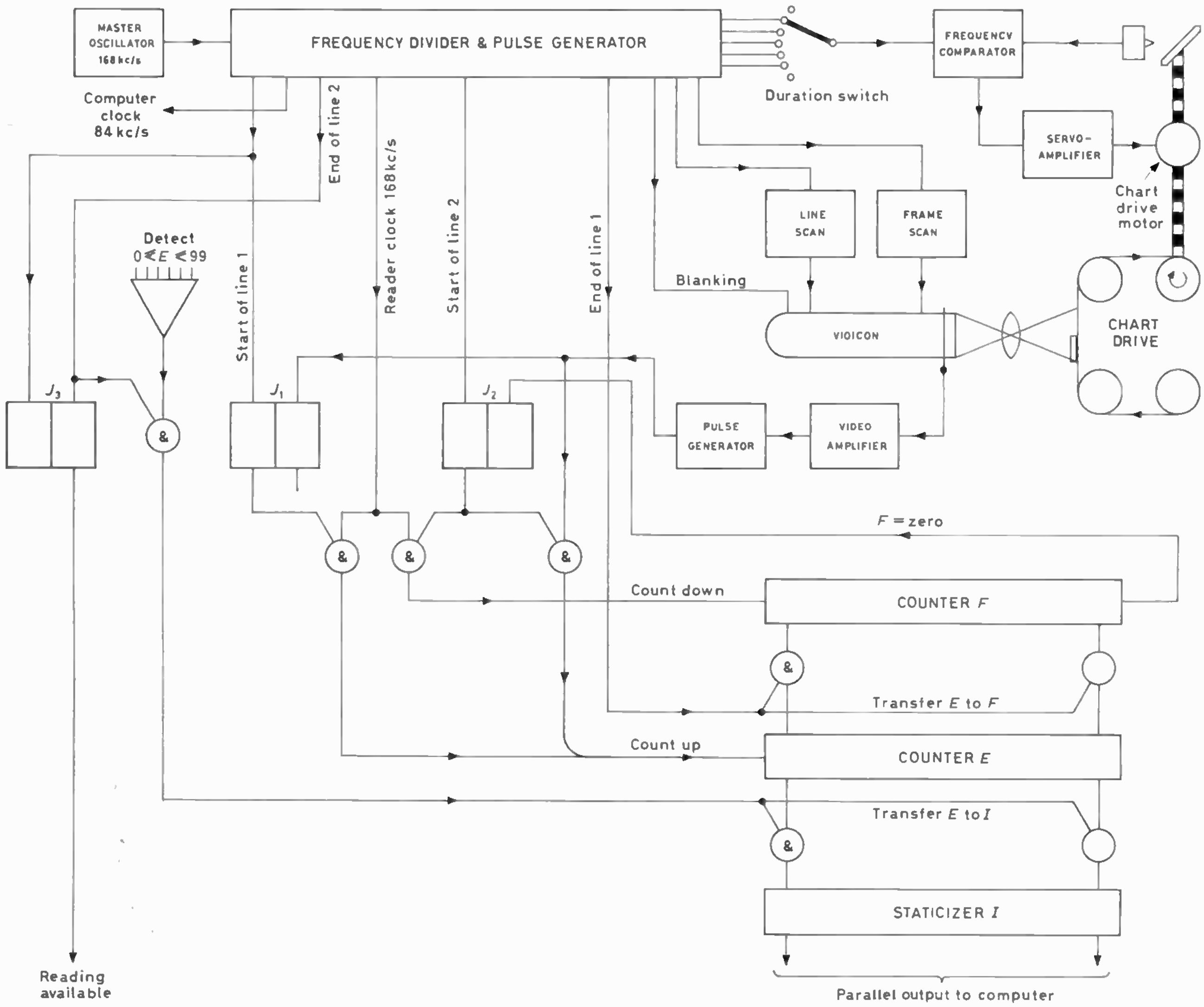



Fig. 4. Reader logic

**The Reader Logic**

Fig. 4 shows the reader logic. The divider chain is fed from a 168kc/s oscillator and produces all the necessary frequencies for the scanning generators and the chart drive velodyne. It also produces a blanking signal for the vidicon and pulses to indicate the start and finish of the two scan lines. Counter $E$ is arranged to count up and its contents can be transferred in parallel to either counter $F$, which counts down, or to staticizer $I$ which feeds the computer. Flip-flops $J_1$ and $J_2$ control the operation of the reader logic.

$J_1$ is set 'on' at the start of the first line and causes counter $E$ to count the 168kc/s clock pulses. This process stops when $J_1$ is set 'off' by the pulse due to the trace on the chart. A pulse occurring at the end of the first line transfers the contents of $E$ to $F$ and resets $E$. When $J_2$ is set 'on' at the start of line 2 register $F$ counts the 168kc/s clock and register $E$ counts the pulses derived from the black bars. This process stops when $J_2$ is set 'off' by $F$ reaching zero.

At this time $E$ holds the desired reading and this is transferred to staticizer $I$ at the end of line 2 provided it lies within the limits 0 to 99.

Flip-flop $J_3$ is always set on at the end of line 2 and indicates to the computer that a new reading is available. It is set off by the computer when it accepts a reading or by the next line 1 pulse. The setting 'on' of $J_3$ is not conditional upon the reading being in range, consequently when an out-of-range reading occurs the previous reading will be repeated to the computer.

The computer works in pure binary and the output is converted to decimal for display purposes. The program is 'wired-in'.

The word length of the computer is either 10 or 13 binary digits depending on the number of readings taken. One word can thus represent 8 191 or 1 923 and larger numbers than this are handled as two words. The arithmetic operations are serial and provision is made for propagating 'carries' from the least significant to the most significant word when adding double length numbers.

The store has a capacity of 64 words of 13 bits. Locations 0 to 49 store the histogram and the remainder are used as working stores, some containing double length numbers.

To derive the mean of a set of readings all the readings are added together and the total is divided by the number of readings. The variance is given to sufficient accuracy by the summation:

$$1/N \sum_{r=0}^{r=49} C_r (r - m)^2$$

where

$C_r$ = number of readings in cell $r$ of histogram

$m$ = address of cell containing the mean

$n$ = total number of readings.

Both mean and variance require double-length working and both require a final division by the number of readings. The capacity of one word is made equal to the number of readings so that the most significant word of a double-length number is automatically the required quotient. Apart from the above, the computations are arranged so as to require only the addition of two numbers, division by 2 and multiplication by 4. An effective subtraction of 1 is achieved by adding a number consisting of all 1's and discarding the carry produced at the most significant end.

The division and multiplication are performed by shifting the number left or right one or two places as appropriate. By deriving variance instead of standard deviation the need for square-rooting is avoided.

The binary-to-decimal conversion is performed serially. The binary number is counted down to zero and simultaneously a decimal counter is counted up from zero. Each decade of this register is connected through weighted registers to moving-coil projection indicators.

### The Computer Programme

The operation is provided with 'clear', 'duration' and 'start' switches to control the computer.

The 'clear' switch sets the computer in a state ready to begin reading a chart. The 'duration' switch selects the correct reference frequency for the chart drive velodyne and the appropriate word length in the computer.

When the 'start' switch is pressed the computer starts the chart-drive motor and performs a routine to clear the core store. This is followed by the actual reading routine in which a reading is accepted from the reader and added to the cumulative total held in the store. The reading is also divided by 2 and used to address the store. This selects the appropriate cell of the histogram and 1 is added to it. Then a test is made to determine if the required number of readings has been received and the routine is repeated if necessary. Finally the computer stops the chart-drive motor, converts the mean from binary to decimal and itself stops with the mean displayed.

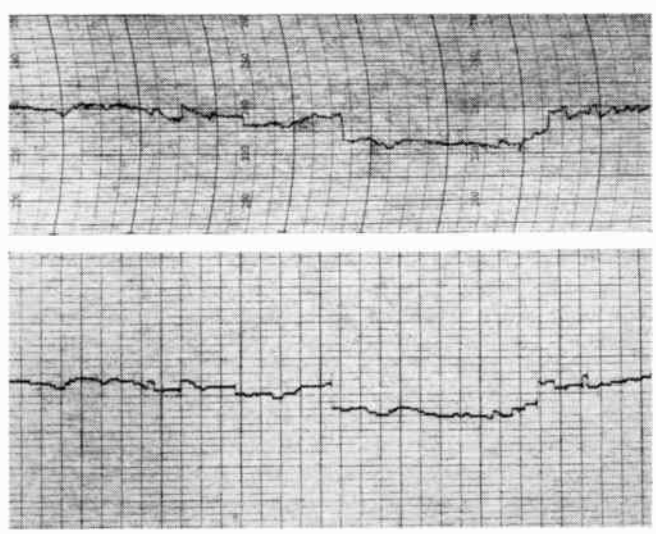
A further operation of the 'start' switch causes the computer to calculate the variance. To do this the contents of each cell in the histogram must be multiplied by the square of the distance of that cell from the mean. This is done by a process of successive addition starting from each end of the histogram in turn and working towards but not including the mean. Allowing for double-length working, this involves over 1 000 additions and takes about 0·2sec. It is followed by binary-to-decimal conversion of the variance and the computer then stops with the variance displayed.

The last phase of the programme enables the histogram to be obtained. Each operation of the 'start' switch causes the contents of the cell specified by two switches on the control panel to be displayed. This part of the programme repeats indefinitely until the 'clear' switch is operated.

### Construction

The computer is accommodated on 70 printed circuit cards each measuring 4½ by 8½in. The reader logic occupies 19 cards and the core store 17 cards. Mullard Ltd

Fig. 5. Comparison of reader output and original chart



Combi-Elements are used for the flip-flops, invertor-amplifiers and the core-store drive and read-out circuits.

**Performance**

The results obtained with the equipment agree closely with manual analysis. With the particular charts used there is a tendency for the variance to be slightly higher than the manual result. This is because the reader picks up large but relatively short-duration excursions of the trace which are ignored with manual analyses. In these cases the analyser is giving a more accurate result. Fig. 5 shows a comparison between the original chart and a copy chart made by converting the digital signal back to analogue and recording it with a high-speed recorder.

The vidicon and its associated circuits are stable enough to operate for long periods without adjustment. The illumination level is so uncritical that the lens may be closed several stops without loss of reading or alternatively sunlight may be superimposed over the artificial lighting of the chart.

The fraction of readings lost due to the obliquity effect is typically about 2 per cent and depends on the nature of the recording.

**Acknowledgments**

The author wishes to thank Dr. I. R. Young of the Naval Division of Evershed & Vignoles Ltd for many helpful discussions, Mr. L. B. S. Golds, M.I.E.E., of the Eastern Electricity Board, who specified the equipment described, and the Directors of Evershed & Vignoles Ltd for permission to publish this article.

# A Transistor Circuit for Xenon Flash Tube Operation

By P. G. M. Dawe*, M.A., M.Sc.(Eng.)

*A circuit is described which lights a xenon flash tube from a transistor controlled switch, operated by pulses obtained from a flip-flop and multivibrator. A feature of the transistor control is the application of an auxiliary ionizing circuit to the lamp to maintain a continuous discharge path within it.*

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

XENON flash tubes have been widely used in commercial stroboscopes[1] and photoflash equipment[2] to provide light flashes of high intensity and short duration, so as to provide a light source for examining or photographing fast moving machinery. The light emitted by this type of xenon source is a continuum built up from a large number of spectral lines, producing a colour approximating to noon sunlight, or white light. The light flash is normally produced by the complete discharge of the stored energy from a large capacitor through a xenon tube, and this action is triggered by the application of a high voltage pulse to a trigger electrode on the tube, the trigger pulse being derived from an auxiliary *LC* circuit (see Fig. 1).

A disadvantage of this type of circuit is that it cannot be operated at high repetition rates, since the capacitance then needs to be greatly reduced for it must always be charged through a resistance that is many times greater than that of the discharge path through the tube and its associated electrodes, in order to protect the power supply. The time-constant of the charging circuit can only be reduced by reducing the value of the capacitance, unless special circuits are adopted to charge it from a much higher voltage, constant current source[3]. Unless such methods are adopted, the energy per flash available at the high rates of operation is low, e.g. less than 0·05 joule for a nominal 10 joule flash lamp at 250 flashes/sec, and the light output falls.

It seemed likely that a transistor circuit might offer a novel method of solving this problem, if used to gate the current through the flash tube, in the form of short rectangular pulses. Some experiments were made to discover the feasibility of this method, having in mind the need to provide a pre-ionized discharge path through the tube which would pass the high current pulses switched through the transistor. It was found that if an alternating voltage (of about 1kV, at 10kc/s) was applied continuously to the trigger electrode of a xenon flash tube, in this instance the F.A. 10, Mazda tube, the gas in the tube would remain sufficiently ionized to maintain a small discharge current (of the order of 30 to 40mA) through it. This current was obtained from a simple 300V d.c. supply connected in series with a high resistance directly across the tube terminals (Fig. 2). It was then found that for a discharge current of 40mA, representing less than two watts of energy continuously dissipated in the tube, the lamp would remain operational to pass much larger currents, of the order of 10A, and so to provide a considerable light output. The larger current was obtained from a 100V source, and this was switched through the tube at first, by a key in series with a resistance, and later by operating a Texas 2G222 transistor with a somewhat lower voltage, this

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being switched by applying a voltage to the base electrode (Fig. 3).

The xenon lamp was envisaged to be finally suitable for operation, for example, in a projector, provided the repetition rate could be made suitably high and the only limitation here would appear to be transistor switching time. In order to examine this possibility a transistor multivibrator circuit[4], mounted on Veroboard, was coupled to a transistor flip-flop circuit[5], similarly mounted, the output of the combination providing pulses which could be varied in width and repetition rate. A typical operation during early experiments made use of pulses of 250$\mu$sec in width and at a repetition rate of 1kc/s. The output from the multivibrator was followed by two stages of emitter-followers, using ACY17 and OC35 transistors.

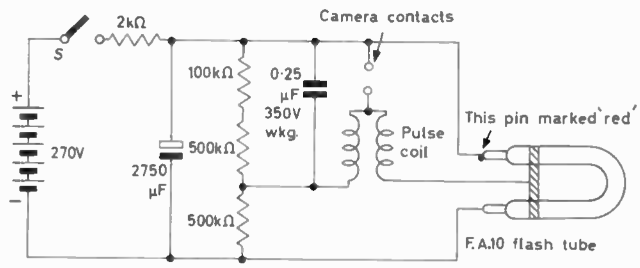


**Fig. 1. Conventional circuit**

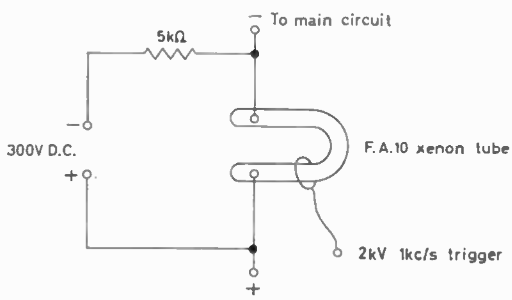


**Fig. 2. Supply of small maintaining current**

These stages were followed by the final current amplifier, a Texas 2S0326 transistor, which was mounted on a 6in by 6in heat sink. The use of a pnp output stage as an alternative, enables negative drive pulses to be used, and thus the emitter-followers are normally off, and this affords a saving in power dissipation. However the 2S0236 npn device was kindly loaned for the experiments by Texas Instruments Ltd and seemed more robust than the pnp devices available at the time, consequently positive drive pulses were employed, and the penultimate OC35 transistor was also provided with a heat sink, 6in square, of ⅛in copper (Fig. 4).

To examine the possibility of using the lamp for projector operation, an F.A. 10 xenon tube was mounted in an Aldis projector. This unit contained a fan for cooling the projection lamp, and it was then found possible to operate the F.A. 10 tube well above the 10W dissipation limit. The optimum operating condition, under forced air cooling, was then as follows:

Operation current: 10A.

Operating voltage: 40 to 45V.

Flash duration: 250$\mu$sec at 1kp/s.

D.C. ionizing current: 40mA.

The flash energy is thus given as:

$$10 \times 40 \times 250 \times 10^{-6} = 0{\cdot}1 \text{ joule/flash}$$

and the continuous power rating:

$$10 \times 40 \times \frac{250}{1\,000} = 100\text{W}.$$

Under these conditions the lamp became of a sufficient brightness to light a screen place 5ft away from the Aldis projector. It was observed that the arc discharge through the lamp, for high currents, of the order of 5 to 10A, was not a completely steady discharge, but was accompanied by a flickering or fluttering effect. It was thought

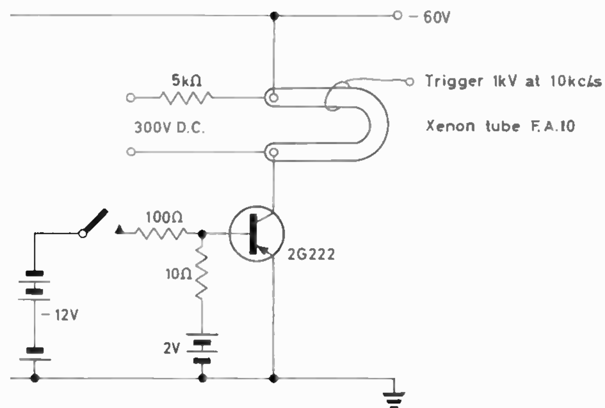


**Fig. 3. Transistor keying arrangement**

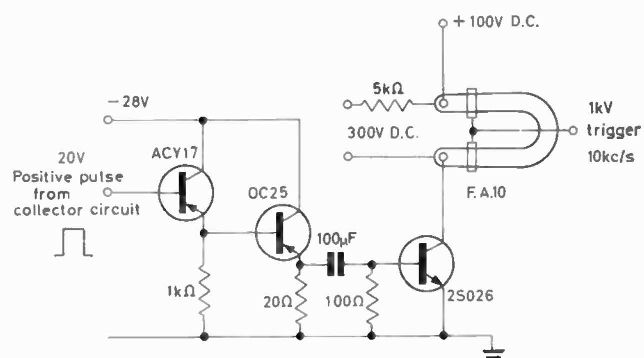


**Fig. 4. Practical circuit arrangement**

that this would be removed if the arc could be made to conduct along the walls of the tube, and it was observed that this could be partly achieved by allowing the tube to get hot. Otherwise a magnetic field might be employed to concentrate the arc, and some experiments were made on these lines.


**Acknowledgments**

Grateful acknowledgment is made to the staff of the Institute of Experimental Psychology for their assistance and to Mr. E. R. F. W. Crossman and Professor R. C. Oldfield for the technical assistance and facilities provided.


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# Frequency Scanning Aerials

By M. F. Radford*, M.A.

*A frequency scanning aerial is an aerial which is so designed that the direction of the radiated beam is a function of frequency. In this article the general principles of such aerials are set forth and it is shown that they may prove an economical alternative to mechanically nodded aerials.*

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

A FREQUENCY scanning aerial is an aerial so designed that the direction of the radiated beam is a function of the signal frequency. By varying the transmitter frequency inertialess beam scanning may be achieved, and since the variation may be introduced electronically in the transmitter drive, the system may be exceedingly agile and versatile.

All frequency scanning aerials contain elements having frequency-sensitive phase characteristics. Two types of aerial are possible, parallel fed and series fed. Parallel fed aerials divide the transmitted power between a number of separate parallel paths, each of which has a different phase/frequency sensitivity and feeds a different part of the aperture (Fig. 1). Series fed aerials use a single long frequency-sensitive line and tap off power at regular intervals to feed the radiators (Fig. 2). The latter system is generally more convenient in practice, as the total amount of path required is much less.

The simplest form of frequency-sensitive element is a length of transmission line, e.g. coaxial line or waveguide. The total phase-shift $\phi$ in a line of length $l$ at wavelength $\lambda$ is given by:

$$\phi = 2\pi l/\lambda_g \quad \text{.................} \quad (1)$$

where for coaxial line $\lambda_g = \lambda$ and for rectangular waveguide:

$$1/\lambda^2 = 1/\lambda_g^2 + 1/(2a)^2 \quad \text{.........} \quad (2)$$

where $a$ is the broad dimension of the guide.

The above equations are strictly true only in vacuo, but for all practical purposes the refractive index of air may be taken as unity.

Frequency sensitivity may be increased by filling the line with some suitable low loss dielectric, or by loading the line with suitable obstacles. Resonant elements such as slotted plates or quarter wave shunt stubs may be used, or alternatively non-resonant elements such as irises or posts may be spaced at resonant intervals. The latter may also be regarded as a line of cascaded resonant cavities. Another way of making a compact dispersive structure which avoids the complications of loading is to use line of small cross-section, e.g. stripline, small coaxial line, ridge or reduced height waveguide.

These highly dispersive compact configurations have two serious disadvantages, high losses and limited power handling capacity. The high losses are caused by higher current densities, particularly in the resonant elements which store considerable amounts of energy. Consequently aerial gain is reduced, and effective cooling systems are required to maintain a constant safe temperature. Power is limited by voltage breakdown, which again is most likely to occur near resonant elements.

A series fed aerial requires only one dispersive line, and this is usually of reasonably large cross-section, no loading being permissible since the line has to handle the full power of the transmitter. Because of the large area and absence of resonances the losses are low. Parallel fed aerials are forced to use compact structures as many dispersive paths are required. The lower power capacity does not matter as each line carries only a small proportion of the total power, but the higher losses are a serious disadvantage. As a result parallel systems are not generally used. However, parallel dispersive lenses may be added to aerials for other purposes, e.g. to correct squint in a linear array, or to remove a frequency-sensitive phase error.

**Radiator Spacing and Angular Coverage**

Whichever system is chosen, it is convenient to use the minimum number of radiators which can achieve the desired scan angles. It will be shown that not all systems are able to scan through the broadside position, i.e. when the beam is exactly perpendicular to the aperture. If this is so the whole scan must lie on one side and a closer radiator spacing is required for a given angular coverage. The radiator spacing is governed by the need to avoid secondary beams, and for an array of infinite aperture the limiting spacing $d$ is given by:

$$d(1 + \sin\theta) = \lambda \quad \text{.............} \quad (3)$$

where $\theta$ is the maximum scan angle (Fig. 3). With this maximum spacing the secondary beam is just emerging from the aperture, and appears at $-90°$ scan if $\theta$ is positive. In practice it is not the peak of the secondary beam but the skirt at a level of 20 to 30dB down that must be suppressed, and for this a slightly closer spacing is required. Shnitkin[1] suggests the formula:

$$d(1 + \sin\theta) = (1 - (1/N))\,\lambda \quad \text{........} \quad (4)$$

where $N$ is the number of uniformly excited elements required to produce an equal beamwidth.

Having determined the radiator spacing, the relative phase shift $\phi$ required between adjacent radiators for a scan angle $\theta$ can now be found. This is given by:

$$\phi = \frac{2\pi d \sin\theta}{\lambda} \quad \text{...............} \quad (5)$$

For a broadside beam of $\theta = 0$, $\phi = 0$ and thus the two radiators must be energized in phase. At centre frequency $f_0$ the phase difference between two adjacent paths in a parallel system or the phase-shift between two adjacent radiator taps in a series system is then a multiple of $2\pi$. In some cases it is more convenient to use an odd multiple of $\pi$ and introduce phase reversals in alternate radiators, thus bringing them all into phase at $f_0$ as before.

The formula (1) may be used to find the length of transmission line required to produce the necessary phase shift with the available frequency excursion. The greater the bandwidth available, the less line will be required and the lower will be the losses.

* *The Marconi Company Ltd.*

Note that the maximum scan angle will not be the same on both sides of the broadside position since the values of $\lambda$ in equation (3) are not equal. Also, the scan rate is non-linear because of the $\sin\theta$ and $\lambda$ terms in equation (5), and possibly because the dispersive line may not have a linear $\phi/f$ relationship.

If $f$ is varied beyond the limit, the secondary beam appears and there will then be two beams so that bearings are ambiguous. If $f$ is varied still more, the original beam vanishes into the array and the secondary beam becomes a new main beam and scans out towards broadside. If the bandwidth of the components is sufficiently great, several beams may be obtained in succession, each sweeping through broadside when the differential phase is some multiple of $2\pi$. This property is useful in some advanced scanning systems. However, in planning aerials using highly dispersive or unusually long lines, care should be taken not to exceed the limits imposed by pulse delay distortion. The latter becomes serious when the energy

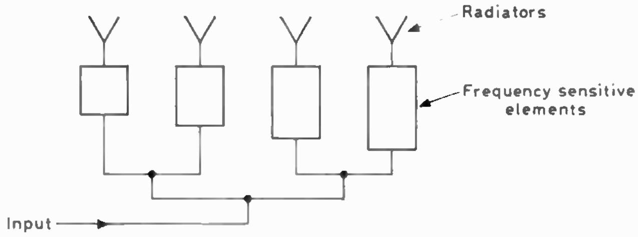



**Fig. 1. Parallel fed array**

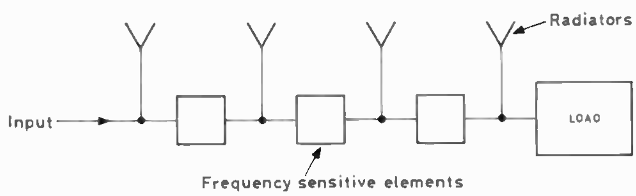



**Fig. 2. Series fed array**

transit time from input to load along a series fed array becomes comparable with the design pulse length.

**Amplitude and Phase Distribution Theory**

It has been shown by J. F. Ramsay[2] that the Fourier transform may be used to calculate radiation patterns from phase and amplitude distributions. In order to obtain a reasonable sidelobe level it is necessary to taper the amplitude distribution towards the ends of the array. In order to scan, a linear phase variation is required, any non-linearity will lead to a deterioration of the beam shape and of the sidelobe level. Typical practical amplitude distributions are:

$$F(x) = 1/5 + 4/5 \cos^3(\pi x/a)$$
$$\text{in the range } (-(a/2) \leqslant x \leqslant (a/2))$$

which gives a 30dB sidelobe level approximately, and:

$$F(x) = 1/7 + 6/7 \cos^2(\pi x/a)$$

which gives a sidelobe level of approximately 35dB. In practice these distributions yield sidelobe levels of 26 to 28dB and 30 to 32dB respectively, allowing for unavoidable small errors. It can be shown that when the number of radiators is greater than about 10, the difference between considering the aperture as a continuous distribution and as an array of point sources is too small to be significant.

It is possible to determine statistically the permissible errors of phase and amplitude at the radiators. If the errors are truly random and the number of radiators is large the tolerances achieved by good mechanical and electrical design are adequate for low sidelobe arrays. If a systematic error occurs, its effect is much more serious, and systematic phase errors of a few degrees may increase sidelobe levels by several decibels. In series fed aerials with tapered amplitude distributions a serious cubic phase error is introduced by the couplers or slots that tap off the power from the dispersive line. These couplers will introduce a phase error in the power coupled out, and also in the power flowing by in the line, the error being a function of the coupling factor, i.e. of the amplitude distribution. The latter error is cumulative and is the chief cause of increased sidelobes.

Two methods of compensation are available, both have been employed in practical arrays. In series compensation the line length between each coupler is adjusted to cancel out the phase error at centre frequency. In parallel compensation simple phase shifters are incorporated in the individual radiator feeds, again set to produce a linear phase front at centre frequency. If the coupling factors

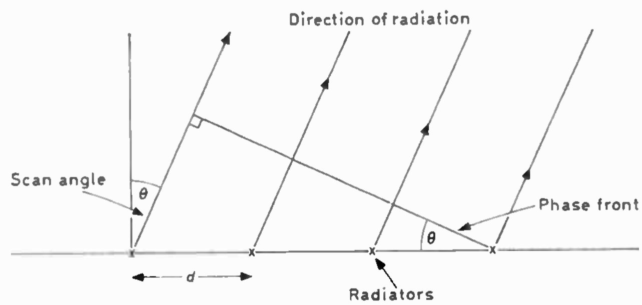



**Fig. 3. Radiator geometry**

are reasonably constant over the working frequency band the correction remains good over a wide scan angle. In certain systems it may be necessary to add slightly dispersive elements to remove additional phase errors towards the ends of the working band.

Amplitude errors are far less troublesome. Small changes in coupling over the band affect the amount of power wasted in the matched termination at the far end of the array, but the slight asymmetry of the distribution has little effect on the radiated pattern.

A practical example of phase correction is shown in Fig. 4. Fig. 4(a) shows the uncorrected radiation pattern of a 12 element experimental series fed array. Fig. 4(b) shows the measured phase error at the aperture and Fig. 4(c) shows the radiation pattern after a set of phase shifters had been added at the outputs of the couplers. The sidelobe level for this array as calculated from the measured amplitude distribution, assuming perfect phase, was 27dB.

**Radiating Elements**

In parallel fed scanning aerials the radiators terminate the dispersive lines, and must therefore be reasonably well matched over the working band. In series fed aerials the radiators have the additional task of coupling the right amount of power out of the feeder, and this can be accomplished in one of two ways. Either the radiator may be directly excited, as, for example, an inclined or displaced slot in a conventional waveguide linear array, or the radiator may be fed from a separate directional coupler. Direct feeding is at first sight simpler, but suffers from a number of serious disadvantages. First, it is difficult to match the radiators perfectly into the line, so that at broad-

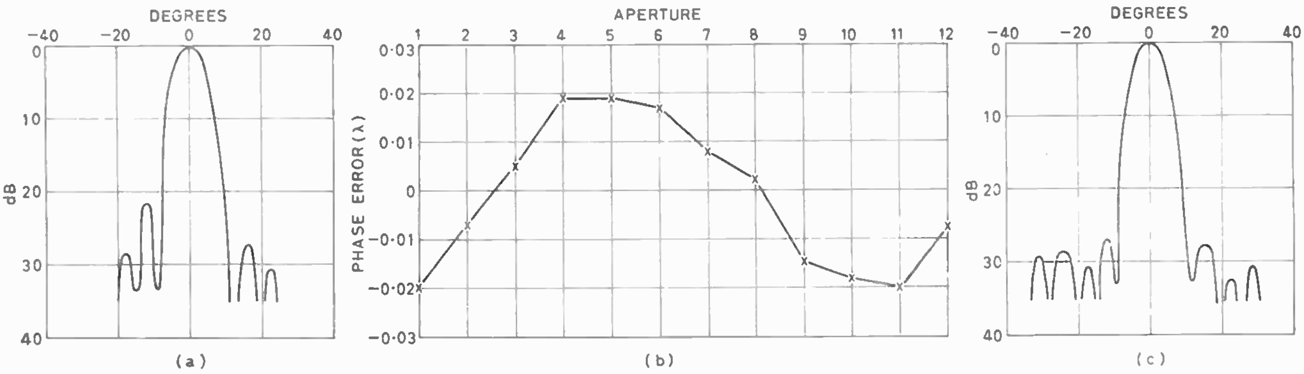



**Fig. 4. Phase correction**

side scan, when separated by a whole number of half-wavelengths, their reflections all add to give a severe mismatch at the transmitter. Secondly mutual interactions between radiators affect the amount of power coupled out, making it very difficult to maintain constant coupling over the band. Even in the absence of mutuals direct fed radiators are not easy to broadband. Thirdly, at certain angles of scan surface waves may form at the aperture and lead to the radiation of large sidelobes. Directional couplers overcome these faults, the load in the fourth arm absorbing most of the spurious energy which would otherwise be reflected or re-radiated. If directional couplers are chosen, the radiator problem becomes identical with that of the parallel fed array.

The radiators are required to have a fairly broad radiation pattern, sufficient to cover the arc of scan without appreciable loss of gain. In consequence, they may be of small size, and so may be closely spaced as required by equation (2). Slots, dipoles, small horns, spirals, helices, and logarithmically periodic structures have all been investigated and there is little difference between their performances, except for the circularly polarized elements which tend to have rather more serious mutual interactions. Logarithmic radiators have low mutuals in linear arrays but have little advantage in two dimensional arrays. In general the choice of radiator is determined by other factors, such as convenience in matching to the transmission line, polarization, and ease of manufacture.

Mutual effects may take the form of directional coupling between adjacent radiators, distorting the radiated field and yet giving very little coupling back into the feeders. This is most noticable with logarithmic or multiple dipole radiators, but occurs to some extent with all types. Since each radiator near the centre of the array has a similar environment, the overall effect on the radiation is negligible over this part of the aperture. Amplitude taper near the edge may lead to small differential variations in the effects of mutuals, this will be unimportant in large arrays where gradients are small. The most serious distribution errors will be at the extremes of the aperture, where radiators may have neighbours on one side only, and these errors can be reduced by including dummy radiators terminated directly in matched loads. In linear arrays one dummy at each end is enough, in planar arrays a row right round the edge may be necessary. However, this precaution should not be taken unless it can be justified experimentally using the particular configuration chosen, as some arrays work well without it while others show a definite improvement after dummies have been added.

It may be necessary to include some form of pressure window or protective radome in the aperture. If so it is important that this should be made an integral part of the design. It must be well matched if it is not to upset the impedance at broadside scan, when all radome reflections add in phase. Also, it must not encourage surface waves. If double skinned, it must be well behaved at various angles of incidence in order to cope with the angular scan. However, a well planned window can contribute to the mechanical strength of the aerial, or it can even be used to trim the input impedance when the main assembly is complete.

**Principles of Design**

From the foregoing it is possible to arrive at the best configuration for any given specification.

In the interests of low losses, simplicity, minimum

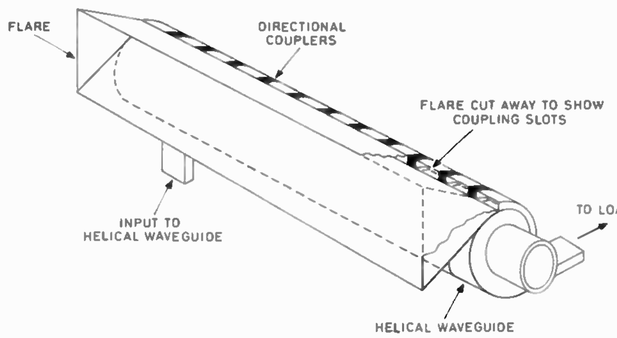
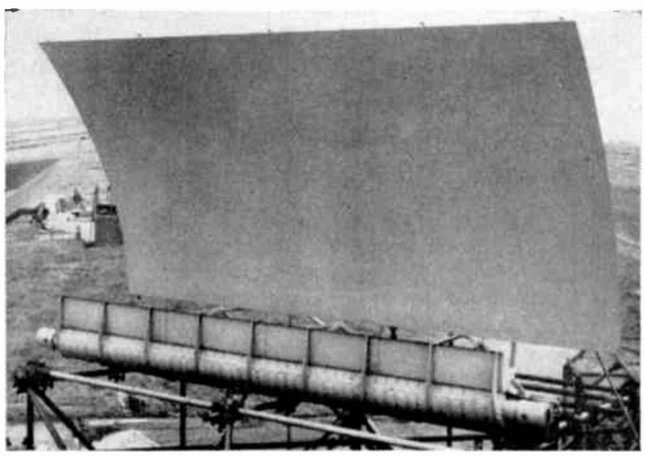



**Fig. 5. Helical waveguide array**

*Helical waveguide antenna on test mount*



weight and bulk the series fed type of aerial is a natural choice in almost every application. Again in the interests of low losses and possibly power capacity, a simple transmission line is almost always better than a loaded line. It will be necessary to fold or coil the line in some manner, and this must be done with the minimum internal reflections. At the same time the line must be suitable for precision manufacture, and must permit output couplers to be added at predetermined spacings to feed the radiators. The only type which is free from reflections is the helical line, and this will usually be a rectangular waveguide (Fig. 5). It can be made very accurately by turning a coarse screw thread on a solid billet and adding an outer wall. The theory of propagation in this guide has been investigated by R. A. Waldron[3].

If many lines are to be packed together to form a planar array or if there is some other severe space limitation, it may be necessary to use a folded line. By making the bends an odd number of quarter wavelengths apart, reflections can be cancelled out almost completely at broadside scan. Phase reversal at alternate radiators is needed since these will now be spaced an odd number of half wavelengths. Limited space and the need for output couplers make rectangular waveguide, even of reduced height, difficult to fit in. Ridge waveguide is better, but the ridge gap is a critical dimension and needs precision constructional techniques. Possibly the best solution is rigid coaxial line of the maximum permissible diameter. The centre conductors then make very convenient vehicles for long multi-hole concentric directional couplers (Fig. 6). It is necessary to develop a suitable 180° bend, and a bend based on a double transition from coaxial line to ridge waveguide[4] has proved satisfactory in experimental arrays.

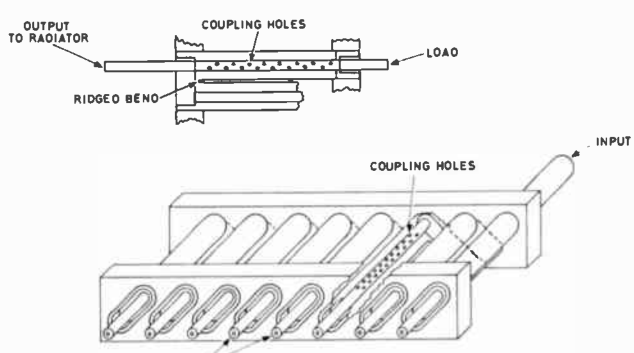
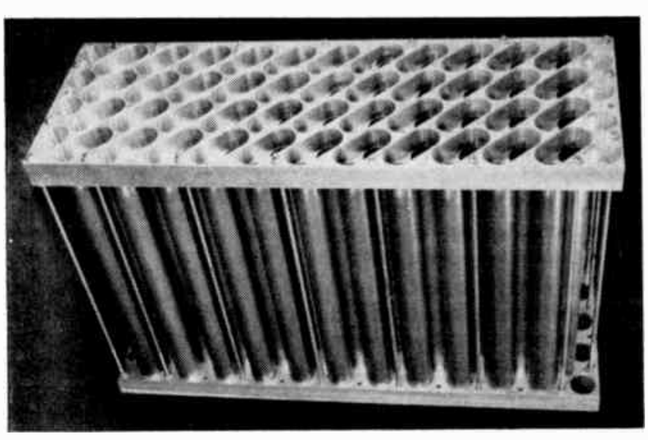



**Fig. 6. Folded coaxial line array**

*Coaxial line array before assembly of inner conductors, and pressings, and radiators*



The advantages of directional couplers have already been stated and it is advisable to use them wherever possible, especially if the frequency bandwidth is large. The coupler match must be first class at the broadside scan frequency, and for this a long coupler is usually most satisfactory. If a helical waveguide is used the coupler must be wound around the outside. Angular guide wavelengths can be matched over a limited range by suitable choice of the coupler waveguide broad dimension and this limits the length of the coupler; however, the limitation does not make the coupler short enough to spoil the design. If the original coupler gives a slight variation of coupling with frequency, misphasing in the curved couplers may be used to obtain a first order compensation.

Phase correction must be added in parallel to the outputs of a helical waveguide, since any modification of the line itself would be exceedingly difficult in practice, although theoretically possible. This can conveniently be carried out by adding short variable width guide sections in the throats of the radiators, in this case horns or waveguides feeding a common flare. Folded lines may be phase corrected by varying the lengths of line between the bends along the array. Once the amount of correction has been established, this is easily incorporated in the manufacture of the folded line.

Apart from ensuring that all critical dimensions are maintained within permissible manufacturing limits, it is necessary to take into account operating temperature when estimating system accuracy. Temperature gradients along the array can introduce phase errors, and as heat dissipation will not be uniform some form of circulating system is required. Liquid or gas cooling may be used, the latter being lighter and cooling the surfaces of the conductors more effectively. The scan angle will vary significantly with the mean temperature of the array, this may be overcome by thermostatic control or by feedback of an angular correction from a thermocouple in the array. In some cases a combination of both methods may be needed.

### Variations of Frequency Scanning

The aerials discussed above provide frequency scanning in one plane only. Any scanning in the orthogonal plane must be provided by some other means. Mechanical rotation is an obvious choice; if this is not fast enough a planar array may be made up of a large number of frequency scanning linear arrays, each fed through a separate phase shifter. Such aerials are complicated and very expensive, and their sidelobe performance is limited by the accuracy of the phase shifters available.

An ingenious form of planar array giving two dimensional coverage by frequency variation alone has been proposed by J. Croney[5]. An array of slightly dispersive arrays is fed from a single highly dispersive array, the latter passing through several modes or 'regimes' in the working frequency band. A coarse raster scan pattern is obtained. By changing the phase of alternate linear arrays by $\pi$ a second interlacing raster may be obtained, giving better coverage and definition.

Another form of scanning which is somewhat akin to frequency scanning has been described by D. E. N. Davies[6]. It is applicable to receiving aerials only. An array of receiving elements feeds a set of identical mixers, one common local oscillator being used. The mixer outputs are combined in a dispersive line, so that when the local oscillator frequency is swept changes appear in the relative phases of the mixer outputs. A fast sawtooth scan may be obtained; however, a separate 'floodlight' transmitting aerial is required.

While two-dimensional electronic scanning has advantages for certain special purposes, it is inevitably more complicated than the straightforward combination of frequency elevation scan and mechanical rotation. The latter provides a sufficiently high data rate for most applications, and yet may prove to be economically competitive with mechanically nodded aerials, since the heavy servomechanisms required to produce a fast vertical scan are eliminated.

**Acknowledgments**

The author wishes to express his thanks to the Board of Admiralty and the Director of Engineering and Research, the Marconi Co. Ltd, for permission to publish this article.

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## Data Link Equipment

It is anticipated that many companies will require data link equipment during the next decade in order to make full use of computer centres. Use of computer centres will become necessary since many companies will not find it very economic to purchase their own computer, yet they will have a very firm requirement for its services in order to reduce overheads, improve overall efficiency and remain competitive. Large companies which already possess a central computer could employ data link equipment to enable their subsidiary offices or factories to make use of this central computer.

The commercial and industrial data link systems which are now available from Ferranti Ltd achieve high degree of accuracy and reliability and are therefore suited for links with computer installations. The same basic system of operation is employed in the three main versions in the range. These versions are as follows:

(1) A one-way system capable of sending data from A to B.

(2) An either-way system capable of sending data from A to B and from B to A, but not at the same time.

(3) A both-ways system capable of sending data from A to B and from B to A simultaneously.

For this range, the tape reader can read information into the system at any rate up to the chosen maximum of 105 characters per second at 1 000 bauds. This maximum character rate is equivalent to the fastest tape punch available in the medium price range. When improvements in punch tape design occur, or a customer requires a higher speed system, the maximum speed of the data link equipment may easily be increased. Character size may be either 5, 6 or 7 bits, or 8 bits if one is a parity bit.

A high degree of accuracy is achieved because an automatic error detection system is built into the equipment which prevents incorrectly received data from reaching the tape punch. If a received block of data contains an error, it is automatically re-transmitted and the tape punch will not record the data on to the tape until it has been correctly received. If the system unnecessarily repeats a block of data or is interrupted due to any cause, the system automatically hesitates and then continues punching without recording any spurious characters.

The one-way link system and the either-way link system require a two-wire telephone circuit to link the transmitter and the receiver; the circuit may be either a private G.P.O. circuit or part of the public switched network. The both-ways link system requires a four-wire private circuit.

Each system consists of a number of basic units and sections and a system may therefore be tailored to suit the user's requirement. For a both-ways system the external units would be located in a small desk unit. On top of the desk would be the tape reader which feeds the data into the system, the control panel which provides the controller with a means of controlling his equipment, and a telephone switched at the control panel, which enables the controller to speak with his counterpart at the other end of the link. Also on the desk would be the tape punch which punches out the received data.

The cabinet below the tape reader would contain the remaining sections for one end of the system. These sections are as follows: (1) Transmitter section; (2) Transmitter storage section; (3) Modulator; (4) Demodulator; (5) Receiver storage section; (6) Receiver section; (7) Clock unit; (8) Automatic request (ARQ) modulator and demodulator.

The output from the tape reader is applied to the transmitter section where it is processed and applied, in block form, to the transmitter storage section.

The transmitter storage section consists of a matrix of ferrite cores, capable of storing two blocks of data. It adds an identifying number to each block, generates the parity bits which are added to the outgoing data, and under the control of the clock unit produces timing waveforms for the rest of the system. The modulator receives the logical waveform from the storage section and converts it, by means of a phase-reversal modulation, into a signal suitable for transmission over a G.P.O. telephone line.

At the distant receiver, the signal is fed, via a control panel, to the demodulator which changes the received signal back into a digital waveform.

The digital waveform is applied to the receiver section, the storage section and the clock unit. The waveform applied to the clock unit enables this unit to produce pulses of the same frequency and phase as the distant transmitter clock unit. These clock pulses are applied to the storage section where, together with the stored block of data, the transmitted block parity is regenerated and passed to the receiver section.

At the receiver section, the block parity received direct from the demodulator is compared with the regenerated block parity signal obtained from the storage system. If the two signals are the same. the receiver station releases the block of data to the tape punch which punches the information on to paper tape.

If, when the receiver section compares its two input signals, there is a difference between them, the receiver section will not release the block of data to the tape punch, but it will instead send a parity failure signal to an automatic request (ARQ) modulator. The ARQ modulator applies a signal to the transmitter section and this signal is sent, via a modulator, back to the transmitter terminal. When the signal reaches the transmitter, it passes through the control unit and is applied to the transmitter ARQ demodulator. The output from this demodulator is in the form of a signal to the transmitter section which instructs the storage section to re-transmit from its store the block of data which was incorrectly received at the receiver section in the distant receiver.

In the case of the both-ways system, multiplexing of the ARQ and transmitted data signals may be achieved by either the time-multiplexing system, which has just been described, or by frequency-multiplexing. Only frequency-multiplexing is employed for the one-way send and one-way receive systems.

If frequency-multiplexing is employed for the both-ways system, it is possible for the transmitter terminal to transmit to one of two distant terminals and receive data from the other at the same time.

The units located in the desk cabinet consist of printed circuit packages which are plugged into mounting shelves, since this system of design and layout is the most consistent with reliability and ease of servicing.

Maximum permissible loop line transmission time for the systems is 60msec at 1 000 bauds. This transmission time is sufficient to allow the equipment to work over almost all the European and Transatlantic circuits.

# Some Simple Dekatron Coupling Circuits

By A. J. Oxley*, M.A., Ph.D.

*Four coupling circuits are described, with serial and parallel carrying, three being suitable for reversible counting. The maximum counting speeds are quoted for each, and all are smaller than the limit imposed by the Dekatron tube itself. Where this is acceptable, however, it is shown that rather simple coupling circuits can be used.*

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

DOUBLE pulse Dekatron tubes have been described quite frequently[1,2,3], so their construction will be taken for granted and only a brief outline of the method of use will be given here, in terms of type GC10B. Both first and second guides ($g_1$ and $g_2$) are normally biased positively by some 40V, causing the glow to sit on a cathode and thereby raising it (if the cathode load is suitable) to about +40V. A negative pulse of about 100V and 80$\mu$sec duration on $g_1$, followed by a similar overlapping pulse on $g_2$, will make the glow rotate clockwise from cathode $C_n$ to $C_{n+1}$ via $g_1$ and $g_2$, and a similar pulse pair in opposite phase will cause anti-clockwise rotation.

Thus the tube is symmetrical, and will add or subtract according to the relative phase of the guide pulses. The symmetry is often unimportant, most counters having a single input lead and registering the total number of pulses arriving there. Reversible counters have add and subtract inputs and are required to form the algebraic sum, i.e. the total number of add input pulses less the total of subtract pulses. More than one decade will invariably be needed in either case, with suitable coupling circuits to ensure that the tens decade adds one count as the glow in the units decade moves from 9 to 0 (but not 1 to 0), and that one tens count is subtracted when the reverse occurs.

One one-way and two bi-directional coupling circuits will be described in some detail, with notes on their performance. In the first, input pulses are fed to all decades simultaneously, via diode gates controlled by the less significant decades. This allows quite simple coupling circuits, using only passive elements, but entails some loss of speed. The system is adaptable to reversible counting.

The other two reversible circuits are designed on the unit principle, each decade unit having two inputs (add and subtract) and two outputs: in a counter, each receives pulses only from the next lower identical decade and sends carry pulses to that next above. They are basically similar, but two cold-cathode gas trigger tubes instead of a vacuum double triode are used per decade in the second, which is suitable only for lower speeds.

Finally, some other reversible circuits are mentioned briefly and possible counting errors considered.

## A One-Way Coupling Circuit

In a one-way counter each decade above the lowest must receive a carry pulse, causing it to add one count, when the next lower decade changes from 9 to 0. Two principles may be used: in the first, the rise of potential at $C_0$ (or fall at $C_9$) energizes a driving circuit which produces guide pulses at the next higher decade. In the second, every counting pulse is fed to all the decades via gates, which admit the pulses only when all lower decades are at 9. For example, if a counter using this system held 19699, the gates for the hundreds and lower decades would be open but for the thousands and above closed. So the next input pulse would advance the three lower decades, giving 19700 as required.

Fig. 1 shows in outline a four-decade counter based on the second principle, using diode gates and GC10B Dekatrons: $g_{13}$ and $C_{93}$, for example, mean the first guides and number 9 cathode of the third (hundreds) decade. It will be assumed that the diodes have zero forward resistance and those in the line AE have back resistance $R$, while the others in combination with parallel resistors have overall back resistance $r'$. The guide bias of 40V (plus some self-bias) and cathode resistors $r$ (150k$\Omega$) cause glowing cathodes to rise 40V above earth.

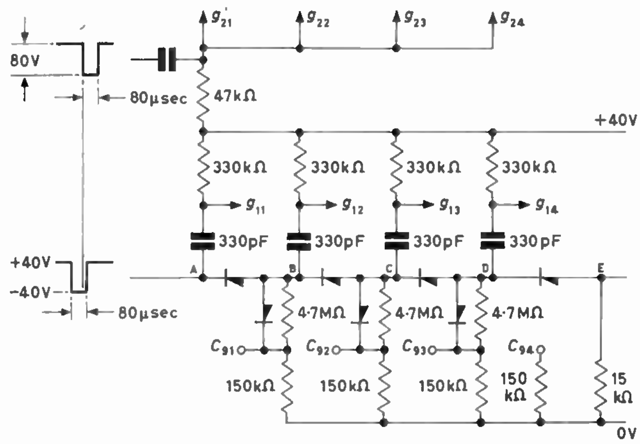


**Fig. 1. Outline of a one-way coupling circuit**

From external circuits, slightly overlapping 80$\mu$sec 80V negative pulses will be applied to A and to the common second guides for each input pulse to the counter. Suppose it holds 9699: then $C_{91}$, $C_{92}$ and $C_{94}$ will be at +40V and $C_{93}$ at zero, making A, B and C +40V and D zero. So when the next input pulse takes A (and B–E) to −40V, 80V pulses will be fed to $g_{11}$, $g_{12}$, and $g_{13}$ but only 40V to $g_{14}$. In consequence the glow will advance in the three lower decades but in the fourth will stay on $C_9$ until the $g_2$ pulse arrives, move to $g_2$, and back to $C_9$ as the pulse dies away. The new count will be 9700.

If it is assumed, say, that a pulse of 61V fed to $g_1$ followed by 80V on $g_2$ will advance the glow, but that one of 59V is not quite sufficient, then any pulse of the right length between 61V and 99V in magnitude at A should operate the counter satisfactorily. In practice, as will be seen, the margin is not so wide partly because of tolerances on the Dekatrons. But in addition the values of $R$ and $r'$ influence the operating range. For example, the upper level of C, achieved with both B and $C_{92}$ at 40V, will be degraded if D is at 0. The potential of $C$ will then be $v$ where:

$$v/(R + r) = (40 - v)/R + (40 - v)/r' \quad \ldots\ldots\ldots (1)$$

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The worst degradation of the lower level will affect B when the glow in all the higher decades is on $C_9$ (e.g. when the counter holds 9992). Where $N$ is the number of decades and $v'$ the potential at B, then:

$$(N-2)(40-v')/r' + (40-v')/R = v'/r \quad \ldots . \quad (2)$$

So $v$ and $v'$ will be close to the ideal values (40 and 0) if $R \gg r' \gg r$. The Dekatron characteristics fix $r$ (at 150kΩ for type GC10B) and $R$ can be made very large (>500MΩ) by using silicon diodes such as the OA202. With $r'$ at 4·7MΩ and $R$ at 500MΩ, $v$ becomes 39·6V and $v' = 2{\cdot}4$V for $N = 4$, but rises to 9V if $r' = 1$MΩ. It is best to use silicon diodes, with a high back resistance at all normal temperatures: the germanium OA90, for example, would be quite useless ($r' \simeq 100$kΩ apart from any resistor in parallel). Too large a value of $r'$ restricts the maximum speed.

circuits at E some gain in speed is achieved by replacing the 4·7MΩ resistors of Fig. 1 by, say, 10MΩ to +150V from B, C and D. Calculations of the maximum speed and degradation of $v$ and $v'$ levels can be carried out in a similar way to those already done.

### Test Performance

Fig. 2 shows the circuit built to test the system described in this section. With no connexions beyond D, larger $g_1$ capacitors, and 10MΩ resistors from B-D to +150V, it was not optimized either for speed or pulse size margin ($v = 40$V, $v' = 7$V). Because they were available, four type GS10C 4kc/s Dekatrons of uncertain history were used, at least one being a reject from another piece of equipment. The input pulses were 200µsec long, of a size sufficient to take $g_2$ from +40V to −40V, and the size of the pulses fed to A was varied by the potentiometer $RV$.

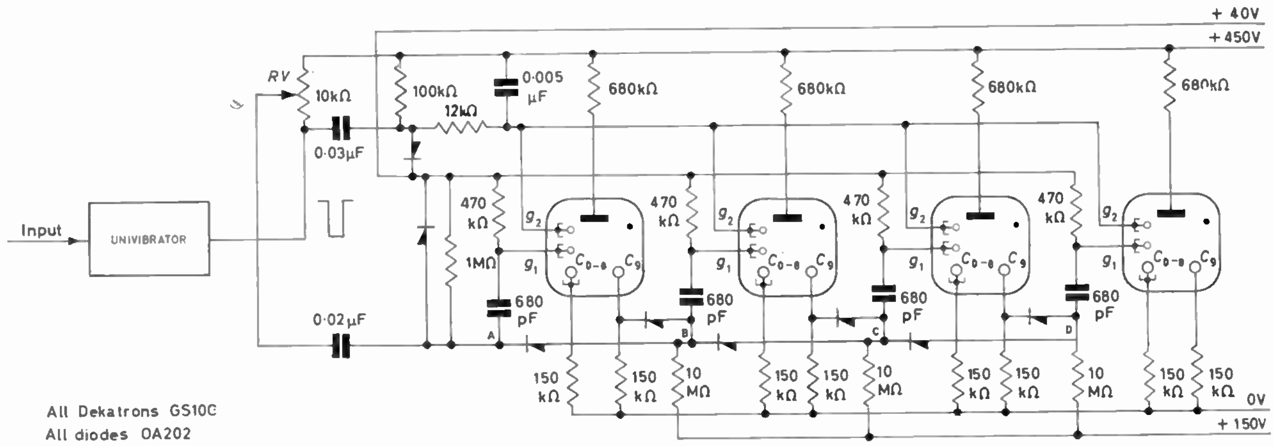


Fig. 2. A four-decade one-way counter

### Operating Speed

After falling to −40V in response to a pulse at A, the points B-D rise quickly to 0V because of the 15kΩ resistor at E. But any rise to 40V is much slower, with a time-constant 330pF × 5MΩ = 1·65msec: 35V will be reached in about 3·3msec. The maximum operating speed will depend on the pulse size margin considered acceptable, but will be around 400c/s. So the system could be used for all decades above the first in a counter using GC10B Dekatrons, of maximum speed 4kc/s.

The speed can be considerably increased by arranging that each input pulse feeds a short positive pulse to E after the end of the $g_2$ pulse, thereby taking B-D to +40V. Those points which should be at 40V (B and C when the counter holds 9699) stay there, while the others fall rapidly to 0 since they now look forwards through diodes to the number 9 cathodes. Resistance $r'$ can now be chosen for minimum degradation of $v$ and $v'$

Looking in more detail it is seen that E should not rise until about 40µsec after the end of the $g_2$ pulse, or the glow transfer in the Dekatrons may be upset. If the $g_1$ resistors are paralleled by diodes the pulse at E can be very short (otherwise it should be to +60V for 330pF × 330kΩ = 110µsec), and about 200µsec must elapse before the next input pulse arrives at A (330pF × 480kΩ = 160µsec), giving maximum speeds of about 2·5kc/s with diodes on the first guides or 2kc/s without. Note that if faster tubes such as the GS10D (rated for 10kc/s) were used instead of type GC10B, component values would need altering but the fractional loss of speed would be about the same, some 40 per cent.

If one wishes to avoid the slight extra complication of

At input frequencies up to 200c/s counting was satisfactory at settings of $RV$ giving between −24V and −44V minimum values at A, and at 400c/s the range was between −28V and −43V. At a median setting, with A swinging from +40V to −35V, a speed of 440c/s was attained.

This performance seems reasonably in accord with expectations, and suggests that the higher maximum speed and better pulse size margin indicated in discussing the circuit at E should be realized to a large extent. The only precaution taken in the test was to trim all the $C_9$ resistors to give outputs of at least +40V (the static level at A): the diodes then ensure that the upper levels of B-D are all equal.

A few words about the Dekatron tubes should be added. As only one free cathode is needed in each, the simplest type (GC10B) can be used. These are normally described as having cathodes 1-9 connected internally and brought out to one base pin, with $C_0$ brought out to another; but this is only conventional, and for the present purpose the common cathodes are regarded as numbers 0-8, with $C_9$ the separate one. The slight disadvantage that the counter must be reset to 9999 instead of 0000 is easily overcome by arranging that the reset system adds one pulse to the counter after setting 9999 in it.

There is a further point to note, important if one is using, say, a three decade counter to divide by 1000. In such an application one would normally connect to the output cathode of the hundreds decade and expect to see one pulse there for each 1000 input pulses. But in the Fig. 1 circuit, because all input pulses go to $g_2$, pulses will appear at $C_{93}$ for each of 100 successive input pulses, and this will be repeated after a further 900 input pulses.

Thus it might be necessary to couple $C_{93}$ to the next circuit via a network of appropriate time-constant so that only the first of the 100 pulses takes effect. In some circumstances the type of output at $C_{93}$ may be an advantage, and it is beneficial to the Dekatrons, which may have to remain with the glow on one cathode for a considerable time in the high decades of a more normal counter, with consequent deterioration of the tube.

A reversible counter on the same principles is outlined in Fig. 8.

### A Fast Reversible Circuit

The symmetrical reversible circuits of this section and the next can conveniently be considered as made up of two parts: the direction-discriminating element, which produces add carry pulses when the glow falls on $C_0$ from $C_9$ but not from $C_1$ (and likewise when subtracting), and the driving circuit, with add and subtract inputs, which produces appropriately phased guide pulses. Both these parts, believed to be novel, can be used independently: for example, this type of discriminator might be combined

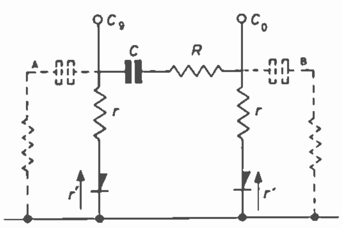


Fig. 3. The basic direction discriminator

with a transistor blocking oscillator drive circuit (Warman and Bibb[4], Bacon[5]) or the drive circuit could be used in conjunction with a Dekatron with 'routing guides' such as the GS10H.

The two parts are not in fact ideally suited to each other, as the discriminator works best with sharp-edged guide pulses, which are not obtainable from the drive circuit. The hard-valve version of the drive circuit, on the other hand, would work rather better with square input grid pulses, which cannot be obtained from the discriminator described. Nevertheless it has been possible to combine these circuit elements to produce one of the simplest of coupling circuits, functioning up to 10kc/s with a GS10D Dekatron but with a rather narrow operating margin.

#### The Direction Discriminator

The basic circuit is shown in Fig. 3, where the diodes are assumed to have resistances of zero forwards and $r'$ backwards. Suppose the Dekatron is adding uniformly, causing the cathode potentials to rise to $V$ as the glow invests them, and that the glow spends time $t_1$ on each cathode and $t_2$ between cathodes. Then as the glow falls on $C_9$ the potential there rises from 0 to $V$, and that at $C_0$ to $Vr/(R + r)$, decaying towards 0 with a time-constant $\tau_1 = C(R + r)$. If $t_1 \gg \tau_1$, $C_0$ will be back at 0 when the next input pulse moves the glow from $C_9$ to $C_0$, but as soon as it leaves $C_9$, $C$ begins to discharge, the discharge current flowing anti-clockwise in Fig. 3. Assuming $r' \gg r$ or $R$, $C_9$ will immediately fall to 0 and $C_0$ to $-V$, decaying with a time-constant $\tau_2 = C(R+2r+r')$. If $\tau_2 \gg t_2$, $C_0$ will rise from $-V$ to $+V$ as the glow invests it, $C_9$ at the same time rising from 0 to $2Vr/(r + R)$ and decaying with time-constant $\tau_1$ towards 0.

In a Dekatron counting at full speed $t_1 = t_2/2$, so by making $C$ small and $r'$ large it is easy to have $\tau_1 \ll t_1$ and $\tau_2 \gg t_2$. Thus when adding, if $R$ and $r$ are set equal, $C_9$ twice jumps positively by $V$ and $C_0$ once by $2V$, after a preliminary negative jump of magnitude $V$. So if a capacitor-resistor combination of time-constant $\ll t_2$ is attached to the cathodes as indicated by the broken lines of Fig. 3, the potential at A will reach $V$ and that at B $2V$ when adding, and vice-versa when the Dekatron is subtracting. Points A and B can then be connected to a suitably-biased drive circuit for the next decade. Notice that $(t_1 + t_2)$ will be larger in a Dekatron counting at low speed (e.g. in the upper decades of a counter), but by using the same drive circuit for all decades $t_2$ is automatically the same for all, so the inequalities above are maintained.

In practice the circuit of Fig. 3 has to be elaborated, and there is no longer such a clear distinction between wanted pulses at B (of height $2V$, when adding) and unwanted (of height $V$, when subtracting). It is often inconvenient to have wanted pulses of duration $\ll t_2$ (i.e. $\ll 67\mu$sec for 10kc/s counting), and the slow rise at the cathode makes such pulses small, so the simple capacitor-resistor coupling to A and B has to be replaced by a capacitor-diode combination (see Fig. 5), to cause fast decay of negative pulses while leaving longer positive pulses. In the original circuit, too, it is essential to have resistors in

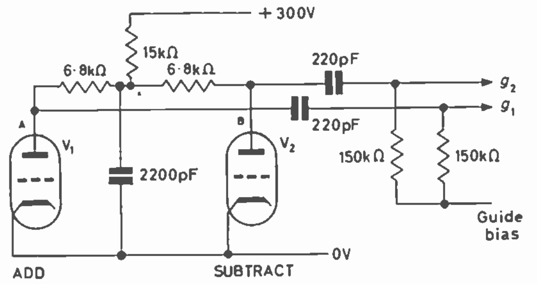


Fig. 4. The basic bi-directional driving circuit

parallel with the (perfect) diodes, because $C_9$ falls to $-V$ as the glow leaves $C_0$ when adding, and it must be back to 0 when the glow falls on $C_9$ a time $(8t_1 + 9t_2)$ later, i.e. one must have $\tau_2 \ll 8(t_1 + t_2)$, or there will be a spurious subtract carry pulse. Remembering that the Dekatron cathodes are high impedance points, these two modifications mean that $C_0$ does not fall to $-V$ as the glow leaves $C_9$, and the difference between wanted and unwanted pulses at B is correspondingly less.

In the full circuit of Fig. 5 it will be seen that resistor $R$ has disappeared and that there are two catching diodes on $C_9$ and $C_0$. Since $C_9$ rises to $2Vr/(r + R)$ as $C_0$ glows when adding, discrimination looks impossible if $R = 0$. But in fact the cathode pulses have long rise-times ($\simeq t_2/2$), especially with the driving circuits used here, so with $\tau_1$ short $C_9$ does not rise very far and $R$ is dispensable. The final glowing cathode potential $V$ is somewhat dependent on the speed at which a given Dekatron is counting, falling by a few volts as the maximum speed is approached. This is highly undesirable, making the unwanted pulse at low speed closer to the wanted pulse at high speed. To avoid this effect, $V$ is stabilized by the catching diodes: the GS10D cathodes rise 30V even at full speed (36V up to 7kc/s, 30V at 12kc/s). The catching diodes are unnecessary if the full speed of the Dekatrons used is not required: when they are included $R$ can be removed whatever the cathode pulse shape.

#### The Driving Circuit

The basic circuit for driving GS10D Dekatrons at up to 10kc/s is shown in Fig. 4. Both halves of the double triode are normally biased beyond cut-off, but when a positive pulse is applied at $V_1$ grid that valve conducts. Anode A falls immediately and B also falls, though not so far and with a delay caused by the 2 200pF capacitor. When the pulse on $V_1$ grid ends and current in $V_1$ ceases, A and B assume equal potentials and both return asymptotically to +300V.

If the changes at A and B were exactly followed by the guides the glow in the Dekatron would not advance, as there would be no incentive for it to move from $g_1$ to $g_2$ (the potential at B is always equal to or above that at A). But when the glow moves to $g_1$ as A falls it begins to charge the 220pF capacitor, so that when A and B become equal at the end of the pulse $g_2$ is lower than $g_1$ and attracts the glow, which subsequently transfers to the next cathode. Oscillation of the glow between $g_2$ and $g_1$ is possible and the component values and grid pulse length must be selected to produce the desired behaviour. The circuit is clearly symmetrical and the glow will move in the opposite direction in response to a pulse on $V_2$ grid.

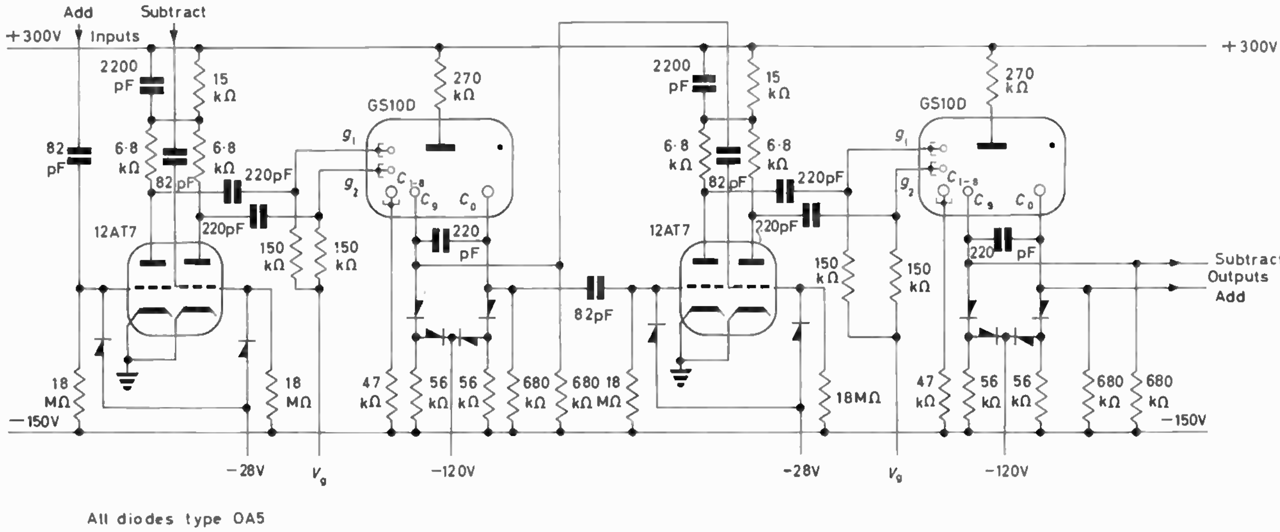


Fig. 5. Two decades of a 10kc/s reversible counter

### The Complete Circuit

Two complete decades are shown in Fig. 5. The grid resistors serve to produce approximately triangular current pulses in the triodes: square waves of fixed duration would probably be better but cannot easily be obtained from the direction discriminator. With the system shown the current flows for a practically fixed time $t_3$ in response to carry pulses or square input pulses longer than $t_3$ fed to the lowest decade. With $t_3$ the same in all decades $t_2$ (actually 30 to 35$\mu$sec) will also be fixed, as required by the direction discriminating circuit. If it were inconvenient to return the 18MΩ grid resistors to −150V, 3·3MΩ to −28V would have a very similar effect.

The guide bias $V_g$ influences the maximum speed attainable. If set at −130V the glow resides mainly on a cathode when the decade is quiescent, whereas at −150V it spreads considerably to the adjacent guides. Below about −140V, apparently harmless oscillations of a few volts appear on the Dekatron electrodes. Maximum speed is achieved at −150V bias, typical test figures being 10kc/s maximum at −130V and 12kc/s at −150V, when the cathodes still reach the −120V level.

To achieve the widest operating margins, thereby allowing for drift of components and tubes, it would be best to set the grid bias on each triode of a multi-decade counter individually, as the tubes are not quite identical. But it is a considerable practical convenience to have a common value, and tests were done this way, using three decades. With $V_g$ at −130V satisfactory test performance was obtained over the grid bias range −24V to −31V at speeds up to 10kc/s, a smaller range than one would like but to be expected because $C_0$ only falls to −163V as the glow leaves $C_9$.

The operating margin would be greater if a lower maximum speed were tolerable, allowing a higher catching level for $C_9$ and $C_0$, or if the triodes were run from a 450V supply to give more gain. Performance figures for a slightly modified counter are quoted in Appendix 2.

Resetting the counter to zero can be achieved as usual by breaking the connexions to all cathodes $C_1$ to $C_9$ or feeding a negative pulse to all cathodes $C_0$, as indicated in Fig. 6. Values of $C$ and $R$ will depend on the number

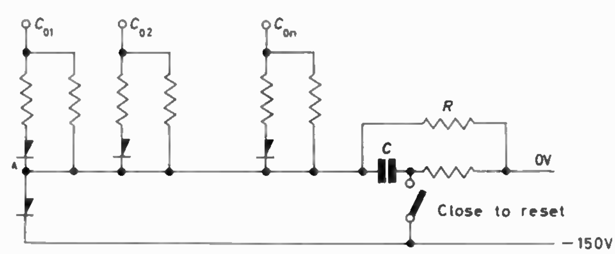


Fig. 6. Reset circuit

of decades in the counter or counters being reset. $C$ must be large enough to make A rise too slowly for carry pulses to be generated immediately after resetting, and $R$ small enough to ensure that A never falls below −150V except when resetting, i.e. not when several cathodes $C_9$ happen to go dark simultaneously.

The decades tested were built with 5 to 10 per cent tolerance resistors and 10 per cent capacitors. Observations during tests indicated the values were not very critical except perhaps for the Dekatron anode resistors, but they should be fairly well matched, especially in critical parts of the discriminator circuit such as the cathode to grid capacitors. The useable grid bias range and maximum speeds do not change much if these capacitors are changed from 68pF to 120pF, but the middle of the range moves, so if a common bias line is to be used it is best to use close tolerance components.

In coupling decade units together it is desirable to keep connecting leads short and avoid stray capacitance wherever possible, and if a bi-directional unit is ever used

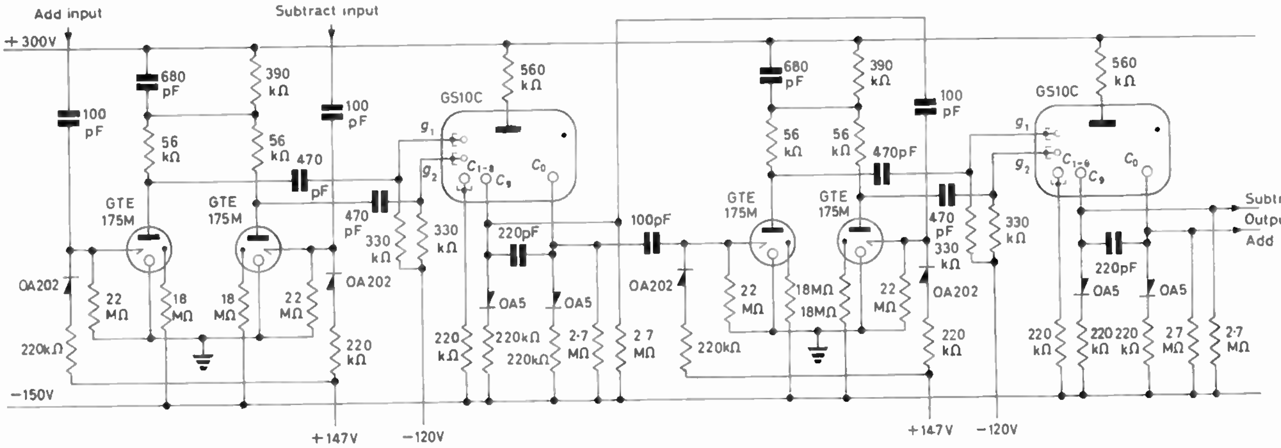


Fig. 7. Two decades of a 500c/s reversible counter

in a one-way counter it should have a dummy load (100kΩ) on the free input terminal. The leading edge of guide pulses to GS10D tubes must not fall by more than 150V/μsec according to published data, so the rise-time of input pulses to the lowest decade must be limited appropriately.

### A Reversible Circuit using Gas Trigger Tubes

This is similar in principle to the circuit of the previous section, but gas trigger tubes type GTE175M have replaced the vacuum triodes. With the limited speed of these trigger tubes the slower GC10/4B or GS10C Dekatrons, rated up to 4kc/s, are more than adequate. Much faster cold-cathode trigger tubes exist, and with appropriate changes of component values the S.T.C. type G1/371K might be used to drive the GS10D Dekatron at its maximum speed of 10kc/s.

The full circuit is shown in Fig. 7. The direction discriminator no longer needs catching diodes on $C_9$ and $C_0$ because the speed of this circuit is so low that the full cathode output is always achieved: it is perhaps worth trimming the cathode resistors to ensure equal signals at all output cathodes of a counter despite slight tube differences. The trigger tubes are self-quenching as the anode resistors are too high to maintain a static discharge, and the 220kΩ trigger resistors are included to avoid excessive trigger current flowing when the tube glows.

In testing a three-decade counter it was found that too low a guide bias caused counter errors at low speed but −120V seemed a safe level and gave a maximum speed of 550c/s to 600c/s with old Dekatrons and 700c/s with new. For both old and new the trigger bias could be varied over the range +140V to +155V before miscounting occurred. The maximum speed falls by about 200c/s at −110V guide bias: at any bias it can probably be improved if the trigger tubes are fed from a +315V supply (the anodes in Fig. 7 fall to about +285V statically because of the current to the auxiliary cathodes). The 390kΩ anode resistor may need to be increased at the 315V level to ensure self-quenching of the trigger tubes, and could be raised to 470kΩ in any case, with only slight loss of speed, as a guarantee against variations of tube characteristics and supply voltage. Stray capacitance should be kept to a minimum in using this circuit.

### Other Reversible Circuits and Dead Time Counting Errors

Fig. 8 shows in outline a circuit intended for reversible counting based on the system described under 'A One-Way Coupling Circuit'. Here a decade should receive the next add input pulse if all lower decades show 9, and the

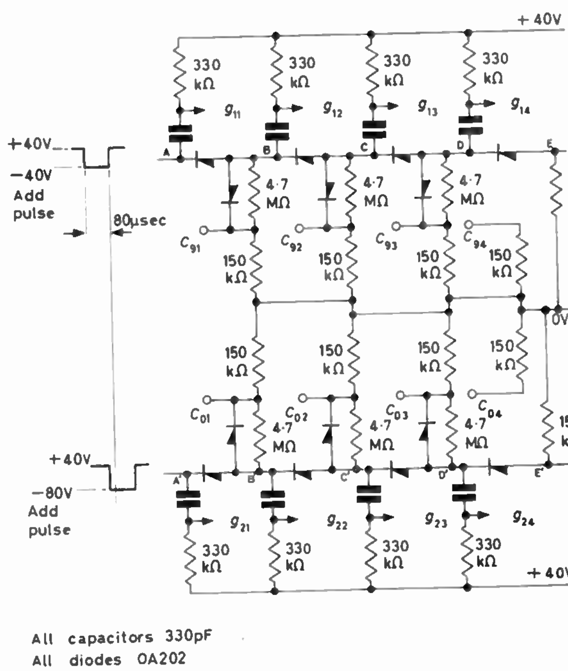


Fig. 8. Outline of a reversible coupling circuit

next subtract input if all lower decades show 0. With the potentials and analysis already given (assuming Dekatrons GS10C, very similar to type GC10B are used), adding is achieved by pulsing A from +40V to −40V and A′ from +40V to −80V after a slight delay, and subtracting by lowering A′ to −40V and A to −80V afterwards. As before, speed is improved by raising E and E′ shortly after each input pulse. This circuit has not been tested, but its performance is likely to be similar to that of the one-way circuit.

Other reversible coupling circuits such as those of Hsu[6] and Oxley[3] have for many purposes been superseded by published circuits[7] for the newer Dekatron types GS10G and GS10H now available, with separately accessible routing guides between $C_9$ and $C_0$. These normally use two double triodes or their equivalent per decade.

Although the maximum speed of a reversible counter is not exceeded errors may occur if an add input pulse follows a subtract input too closely, or in some applications even overlaps it. Possible cases of these awkward inputs have been listed with a circuit adapted to handle most of them

(Oxley[3]). If such inputs do arise, miscounting is likely in the circuits of Figs. 5, 7, and 8, depending on the minimum time between opposite input pulses, and the Dekatrons with routing guides should be used instead, preferably with d.c. coupling between decades.

Prospective counting failures arising from normal inputs due to a slow drift in circuit components can usually be detected, by periodically checking the performance with deliberately off-set pulse sizes in the circuits of Figs. 2 and 8 and off-set bias levels in those of Figs. 5 and 7.

**Acknowledgments**

The author is glad to acknowledge the assistance of A. G. Forson and the Bubble Chamber Group at Cambridge, and the support of the United Kingdom Atomic Energy Authority and Department of Scientific and Industrial Research for parts of the work.

## Appendix

### (1) The Function of the Diodes

The essential function of the diodes in Fig. 3 is to make $\tau_2 \gg \tau_1$, so that even when counting at the maximum speed of the Dekatron (i.e. with $t_2 = 2t_1$) $t_1 \gg \tau_1$ and $\tau_2 \gg t_2$. But in the reversible circuit using gas trigger tubes $t_2 = 150\mu$sec and $t_1 \not< 1250\mu$sec because of the low speed attainable. So it is possible to dispense with the OA5 diodes, at the same time increasing the $C_9$ and $C_0$ resistors and the capacitor between them (to 390kΩ and 1200pF in a test circuit, making $\tau_1$ about 700$\mu$sec and $\tau_2$ 940$\mu$sec). However, the diodes increase the difference between wanted and unwanted signals, and without them the useable trigger bias range in a two-decade counter tested was +150V to +158V, so it is recommended that the diodes be used.

### (2) A Modified Drive Circuit

The fast reversible circuit was designed to fit in with an existing machine. The numbers in the counter were to be punched on paper tape for feeding to a computer, the method being to apply 9 pulses to the add input of each decade in turn and detect how many were needed to move the decade past $C_0$ (see Oxley[8]). The 9 pulses were 20$\mu$sec duration and +10V high, arriving at a repetition frequency of 10kc/s.

The circuit on the grid of the add triode was modified as shown in Fig. 9, a change which added stray capacitance at this high impedance point and inevitably worsened the

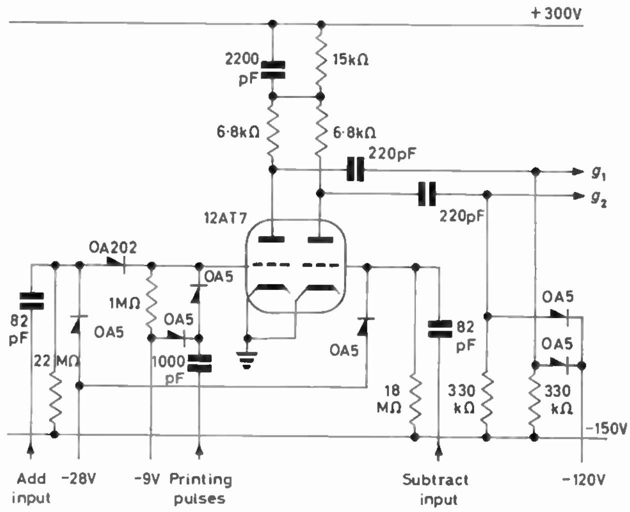


**Fig. 9. Modified drive circuit compatible with printing system**

operating margins, especially in view of the necessity to respond correctly to the 9 small pulses mentioned above. It was found necessary (Forson[9]) to modify the guide circuit as in Fig. 9 for correct performance. With a common grid bias for 8 decade units of this type the operating grid bias range had shrunk to 2V after 9 months' use (Forson[9]). This is unsatisfactory, but the range for normal counting was rather larger and should certainly be greater for straightforward counter decades. Potentiometers to fix the optimum bias for each triode separately would have been an advantage.

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# A New Type of Free-Running Multivibrator

By V. M. Ristic*

*This article describes a series type emitter coupled multivibrator using two pnp transistors. The circuit has good rise and fall times and will oscillate between 0·01c/s and 250kc/s.*

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

The article describes a serial type emitter coupled multivibrator with two pnp transistors.

Many solutions have been proposed for setting the rise and fall times of transistor multivibrators[2-6]. The proposed circuit, Fig. 1, combines a number of useful features.

If the time-constants $R_{b1}C_1$ and $R_{b2}C_2$ are sufficiently large the frequency of oscillation is determined by $CR_1$ and $CR_2$ only. It is interesting to note that the coupling $C_1R_2$ is a Bootstrap one.

When the voltage $-V_{cc}$ is switched on, the emitter of the transistor $VT_1$ is at the potential of point B, the capacitor $C$ is charged via the transistor $VT_1$ which is in the saturated state. The charging current of the capacitor $C$ keeps the transistor $VT_2$ in a non-conductive state. As the current decreases, the transistor $VT_2$ comes out of the non-conducting state and enters a conductive state. The voltage on capacitor $C$ acts now as an equivalent collector

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voltage for transistor $VT_2$. The capacitor $C$ discharges through the resistor $R_2$ and the transistor $VT_2$. When the discharging current decreases $VT_1$ enters a conductive state and the whole cycle is repeated.

### Calculation of the Oscillation Period

Suppose transistor $VT_1$ at time $t = 0$ enters the saturated state. The voltage of capacitor $C$ is $-V_{c1}$ (Fig. 2). The current $i_1$ will equal to:

$$i_1 = \frac{V_{cc} - V_{c1}}{R_1} \exp(-(t/CR_1)) \quad \text{.......... (1)}$$

The voltage across the capacitor $C$ is:

$$u = -(V_{cc} - V_{c1})\,[1 - \exp(-(t/CR_1))] - V_{c1} \quad \text{............ (2)}$$

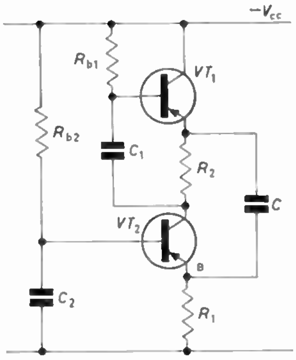


Fig. 1. The multivibrator circuit

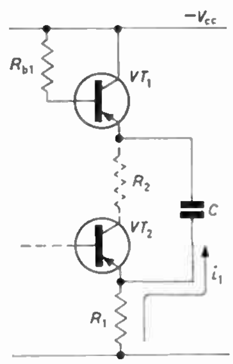


Fig. 2. $VT_1$ conducting

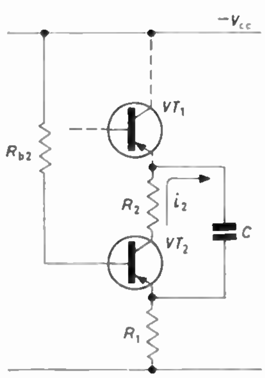


Fig. 3. $VT_2$ conducting

At time $t = T_1$ transistor $VT_1$ reaches a non-conductive region.

Suppose that the voltage of the capacitor $C$ is now $-V_{c2}$. From equation (2) derives:

$$T_1 = CR_1 \ln \frac{V_{cc} - V_{c1}}{V_{cc} - V_{c2}} \quad \text{........... (3)}$$

In the next half period the voltage on capacitor $C$ acts as an equivalent collector voltage (Fig. 3).

The voltage across the capacitor $C$ is:

$$u = -V_{c2} \exp\left(-\frac{t - T_1}{CR_2}\right) \quad \text{............ (4)}$$

For $t = T_1 + T_2$ $u = -V_{c1}$ and from equation (4):

$$T_2 = CR_2 \ln (V_{c2}/V_{c1}) \quad \text{............ (5)}$$

Experimentally it is found that for $R_1 = 7\text{k}\Omega$, $R_2 = 2\text{k}\Omega$, $C = 11\,300\text{pF}$, $V_{cc} = 12\text{V}$, $VT_1$ and $VT_2$, OC45 the voltages $V_{c1}$ and $V_{c2}$ are:

$$V_{c1} = 1\text{V} \quad V_{c2} = 5\text{V} \quad \text{............ (6)}$$

Substituting equation (6) in equations (3) and (5) it is calculated for $T_1 = 30{\cdot}5\mu\text{sec}$ (measured $31\mu$sec) and $T_2 = 36\mu\text{sec}$ (measured $35\mu$sec). Voltage waveforms are given in Fig. 4.

### The Time-Constants $R_{b1}C_1$ and $R_{b2}C_2$

The time-constants $R_{b1}C_1$ and $R_{b2}C_2$ determine the length of the transient state from the moment of switching on the voltage $V_{cc}$ to the moment when the multivibrator enters the stable region of oscillation. Accordingly it is important that these two constants are as small as possible. It is obvious that only $C_1$ and $C_2$ can be changed because $R_{b1}$ and $R_{b2}$ decide the position of the working point when $VT_1$ and $VT_2$ are conductive respectively.

### Experimental Results

As $R_1 = 7\text{k}\Omega$, $R_2 = 2\text{k}\Omega$, $C = 11\,300\text{pF}$, $R_{b1} = 6\text{k}\Omega$, $R_{b2} = 31\text{k}\Omega$ and $C_1 = C_2 = 8\mu\text{F}$, therefore $V_{c1} = 1\text{V}$, $V_{c2} = 5\text{V}$, $T_1 = 31\mu\text{sec}$, $T_2 = 35\mu\text{sec}$ and $\tau_r = 0{\cdot}6\mu\text{sec}$. The circuit can be loaded with $R = 27\text{k}\Omega$ or a capacitive load of 800pF, a series $RC$ load of $15\text{k}\Omega$ and $0{\cdot}1\mu\text{F}$, or a parallel $RC$ load of 500pF and $30\text{k}\Omega$. The circuit works with voltages from 2 to 24V. The frequency of the multivibrator can be changed from 0·01c/s to 250kc/s.

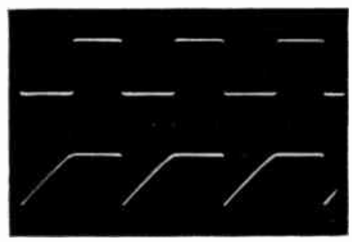

Fig. 4. Collector and emitter voltage waveforms of $VT_2$

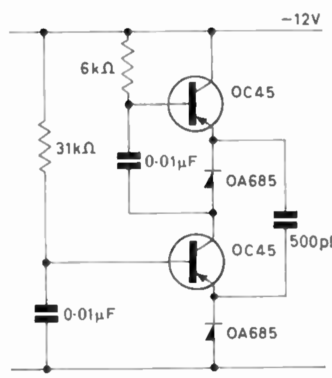


Fig. 5. Circuit for maximum frequency

(600kc/s, rise time 0·2μsec)

By inserting diodes instead of the resistors $R_1$ and $R_2$ one may obtain the maximum oscillating frequency for a given type of transistor (Fig. 5).

### Conclusion

The advantage of the circuit is a good rise and fall time. Power consumption takes place only in one half period (when $VT_1$ conducts). The temperature dependence of the period of oscillation is not critical because of emitter resistances $R_1$ and $R_2$.


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# Gain Controlled Band-Pass Amplifiers

## A New Approach Using Integrated Circuit Video Amplifiers

**(Part 2)**

By M. D. Wood*, M.A.

(Voir page 205 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 213)

THE idea of using a video amplifier of several stages between tuned input and output stages is a very simple one. It is similar to that of using a comparatively broad band tuned i.f. strip whose final band shape is determined by a crystal filter. The reason that this idea is rarely given more than passing consideration is what is called 'gain-bandwidth product'. This can be explained simply by stating that if the gain of an amplifier stage is increased $n$-fold, its bandwidth is reduced $n$-fold. Normally this product is associated with the output circuit, for which there are two relations:

$$\text{Gain} \propto R_L$$

$$\text{Bandwidth} \propto 1/CR_L$$

where $R_L$ is the load resistance and $C$ is the output (plus stray) capacitance shunting $R_L$. The product $G \times B$ is a constant proportional to $1/C$.

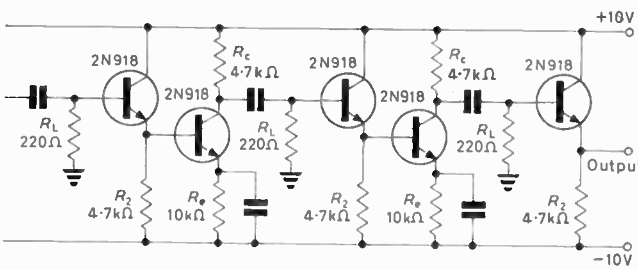


**Fig. 11.** Simple video amplifier

In view of this fact it would obviously seem to be uneconomic to build up a given overall gain using stages with a bandwidth many times that required—less stages of higher gain could be used if their bandwidths were reduced by tuning. The problem is seen to be more one of economics and the number of stages and components, than one of circuit design. However, when considering transistor rather than valve amplifiers, these facts can be considered in a different light.

Firstly, transistors can be considered and used as 'just another component': the fact that one might use an extra transistor and less of other components makes little difference to the price or complexity. Secondly, this being so, there is the big advantage of reducing the number of tuned circuits (with their alignment problems) and replacing them by simple video stages. Thirdly, and associated with this, is the case of design whereby a single circuit configuration can be used for all the video stages in any amplifier at any frequency up to (as will be shown) 100Mc/s or more. Lastly, gain control exercised on or in the video section has no effect on the band shape as determined by the input and output stages.

### Video Amplifier Circuits

Probably the simplest video circuit is that of a transistor with resistive collector load and a bypassed emitter resistor. This must have an emitter-follower drive to provide a low impedance source (without which it is found that there is a slope on the amplitude/frequency response), and it must be followed by an emitter-follower to keep to a minimum the capacitance across the collector load resistance. A typical circuit is shown in Fig. 11. The resistors $R_C$ are a convenient value for the d.c. conditions, the actual (a.c.) load resistor is $R_L$. By using $R_L$ in this manner, and by using both positive and negative supply rails, the circuit is made very simple and uses a minimum of components. The stage gain is given approximately by:

$$G = R_L/r_e$$

If the lower ends of resistors $R_e$ are decoupled and returned to a variable voltage a.g.c. line, the gain can be varied by the current dependence of $r_e$. The gain can be varied from a real gain of about 15dB (depending on the value of $R_L$) to an attenuation of about 20dB. Using transistors type ZT918 (2N918) with a typical $f_1$ of 900Mc/s, a voltage gain of ×4 (12dB) flat to 40Mc/s, −3dB point about 60Mc/s is obtained—or pro-rata, playing off gain versus frequency response.

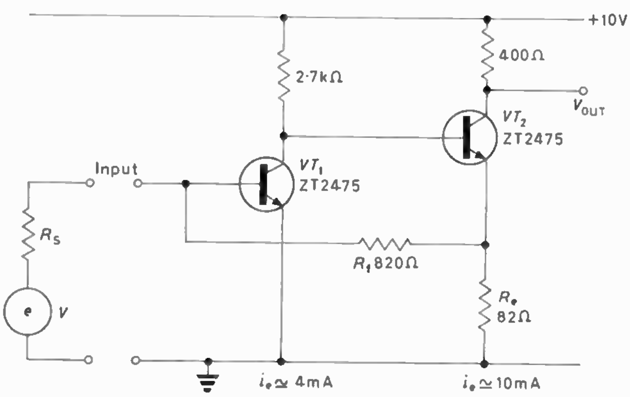


**Fig. 12.** Shunt-series feedback pair video amplifier

While this is not the circuit with the best gain-bandwidth product, it does have a wide gain control range and is very simple. Furthermore, using the word component as above, this circuit uses no more components for a given overall gain than does an amplifier with tuned stages.

Apart from this simple video circuit, there are several circuits using two or more transistors in a feedback loop[13]. One of these, the shunt-series-feedback-pair[13,14] is particularly suitable for the present application. It is this circuit which has been developed in integrated form as device type ZLA10, the circuit diagram of which is shown in Fig. 12.

The current gain of this circuit is theoretically given by:

$$G_i = \frac{R_f + R_e}{R_e} = \times 11$$

but in practice there is a loss of about 10 per cent due to non-infinite $\beta$ in the transistors, i.e. actual gain $\simeq \times 10$ (20dB). The circuit has an input impedance of about 10Ω

* *Ferranti Ltd.*

and an output impedance equal to the 390Ω collector resistance.

The circuit can, alternatively, be considered as a voltage amplifier such that when fed from a source $V$ of internal impedance $R_s$ (Fig. 12).

$$G_v = V_{out}/V = R_f/R_s \times 400/R_e = 4000/R_s$$

(c.f. anode-follower type input circuit). If stages are cascaded so that $R_s = 400\Omega$ = output impedance of previous stage, the voltage gain becomes ×10 (20dB).

The frequency response of a single stage when feeding into a high impedance detector (4pF load) is flat to 85Mc/s with a −3dB point at about 100Mc/s. If the output is a.c. coupled into a 50Ω load, thereby reducing the effect of capacitive shunting of the output, the voltage (but not the current) gain is reduced and the −3dB point is about 200Mc/s.

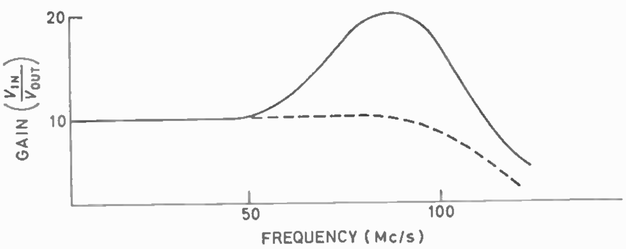


**Fig. 13. Frequency response of feedback pair amplifier**
*Full curve—uncompensated; dashed curve—compensated*

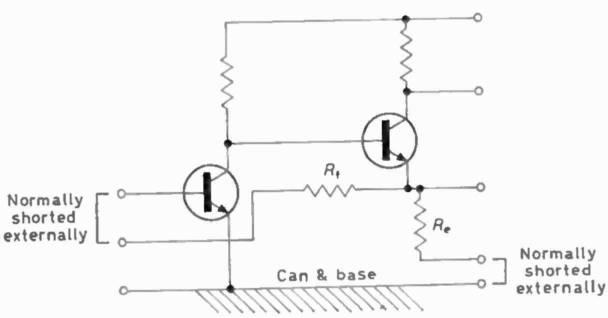


**Fig. 14. External connexions on integrated video amplifier**

## Integrated Amplifier Design

Ghausi[14] gives a very full design treatment of the shunt series feedback pair which is the basis of this amplifier. The most important factor in this treatment is that circuit capacitances and high frequency effects in the transistors cause the gain of the amplifier to rise at high frequencies. At even higher frequencies the gain falls off due to shunt capacitances, giving an overall frequency response as in Fig. 13 (full curve). This can be compensated by shunting $R_f$ in Fig. 12 with a capacitance of about 2pF. This increases the feedback at high frequencies and leads to the dotted curve in Fig. 13.

When the circuit was developed using conventional components it was found that the $R_e$, $R_f$ values could be chosen so that stray capacitance alone across $R_f$ cancelled the peak. However, when the transfer to the integrated silicon resistors was made, it was found that two things had happened. Firstly, while the general level of stray capacitance had decreased giving extended high frequency performance, the capacitance across $R_e$ had not decreased and the h.f. peak reappeared (see later). To correct this it is necessary to add 2·2pF externally across $R_f$, and it is under these conditions that the above quoted frequency responses were measured. The amplifier is mounted in a reduced height eight lead VASCA S0-3, JEDEC T0-5 can with a VASCA SB8-1 base. The various leads are brought out to the pins as shown in Fig. 14. With these connexions it is possible to monitor the important resistor values (all except the 2·7kΩ), and it is also possible to vary the circuit by increasing $R_f$ and $R_e$ externally.

Successive amplifier circuits can be cascaded merely by connecting output to input with a capacitor. The capacitive impedance should be small compared to 10Ω at the lowest frequency of interest. Under these conditions the circuit acts as a true current amplifier, and virtually all of the (a.c.) output current from one stage flows into the low input impedance of the next stage. This low input impedance has, however, much further significance. Because of it, stray capacitance between stages and the output capacitance of any stage are no longer the factors determining the overall bandwidth (c.f. the frequency response quoted earlier when a single stage is loaded with 50Ω). For the amplifier loop itself, i.e. neglecting the effect of capacitance across the output, the −3dB point is of the order of 300Mc/s. If several stages are cascaded, the amplifier bandwidth reduces according to the usual bandwidth shrinkage formula:

$$\beta_o = \beta\sqrt{(2^{1/n} - 1)} \text{ or approx. } \beta_o = \beta/1{\cdot}2\sqrt{n}$$

where $n$ = number of cascaded stages of individual bandwidth $\beta$.

Thus, three stages would halve the bandwidth to about 150Mc/s, and unless a low load impedance were used on the last stage the capacitance across the 400Ω would still be the major bandwidth limitation. Thus four stages have been cascaded to give 68dB gain (see later) with a 3dB point of 90Mc/s when feeding a high impedance detector load.

## Tuned Amplifier Using Video Stages

It is now possible to consider the overall design of a tuned amplifier using video stages. Firstly, it has been seen that the basic video stage in integrated form has a frequency response up to about 300Mc/s if the effect of capacitance across the load is neglected. It is this figure which is to be used as the basis of bandwidth reduction when stages are cascaded. If it can be arranged to terminate the last stage in a very low load resistance, or alternatively, if a tuned circuit can be put at this point to tune out the output capacitance, then there need be no further bandwidth reduction. On this basis, a video gain of 80dB up to 100Mc/s is feasible.

Because the interstage impedance level is very low, stray capacitance at this point is not important. Furthermore, the a.c. voltage level is low, reducing electrostatic radiation, while the low impedance reduces electrostatic pick-up. To reduce electromagnetic radiation and pick-up, the interconnecting lead and capacitor can be screened by or placed alongside, the earth return lead. The complete amplifier element is of course screened by its can. Each stage must be decoupled externally (see later) and normal care has to be taken over the best earth point to which to decouple.

It is difficult to generalize about the types of stage that should be used for the tuned input and output, and how these will be connected to the video stages. This will depend largely on the particular application. In any low level application noise is likely to be important and a low noise input stage will be necessary. The output of this stage should be tuned to reduce the noise bandwidth and prevent cross-modulation by out-of-band noise. The gain of the tuned input stage should be large enough (say 20dB) to make the video stage input noise insignificant. At the other end, the output should also be tuned to eliminate what could otherwise be a comparatively large power from wide-band noise generated in the first video stage. In this respect it is best to use the smallest interstage coupling capacitors possible without infringing on the low end of the required pass band. This also reduces the l.f. decoupling required.

Various tuned circuit arrangements are possible depending on the actual band shape required.

(1) Input and output synchronously tuned.

(2) Input and output stagger tuned as a 'pair'. Not advisable with a large video gain because of the unsymmetrical effects due to noise and overloading.

(3) A double tuned input circuit and centre-tuned output circuit forming a 'triple'.

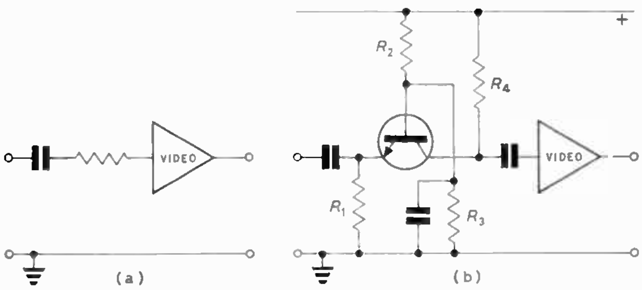


**Fig. 15. Input circuit connexion to video amplifier**
*(a) Series resistor; (b) grounded base*

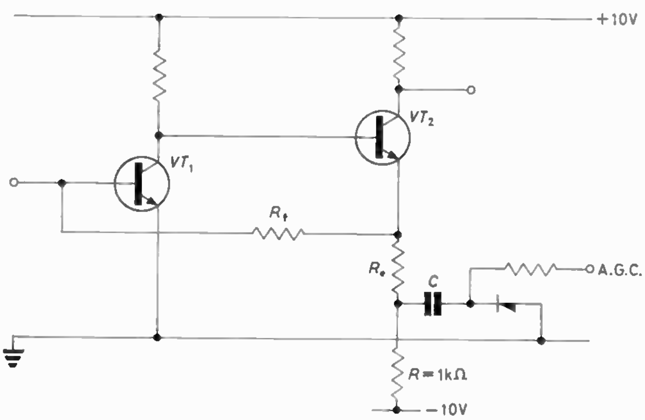


**Fig. 16. A.G.C. applied to feedback pair—variable emitter decoupling**

(4) Double tuned input and output circuits for a wide-band, steep sided response.

While more complicated band shapes could be obtained by adding further tuned input or output stages, a crystal filter would be worth considering as an alternative.

Where the amplifier is to be used between 50 and 100Mc/s it will be necessary to tune the output of the last video stage, although the built-in 400Ω collector load may be rather lower than desirable (see later). At lower frequencies the input capacitance of an emitter-follower is tolerable, and this can drive the required tuned output stage.

The feed to the input video stage presents rather more of a problem because not only is this a very low impedance (10Ω), but by virtue of the feedback loop it is also slightly inductive, causing the input impedance to rise with frequency. Furthermore, the source resistance must be greater than the actual input impedance of the first transistor ($VT_1$ in Fig. 12) otherwise the current fed back through $R_f$ will not flow into the base of $VT_1$, and the loop gain will fall. Two possible connexions are shown in Fig. 15. In Fig. 15(a) a series resistor is used to provide the high source impedance required by the video stage, and some or all of the damping on the tuned circuit feeding it. This is of course a lossy method because signal power is dissipated in $R$. In Fig. 15(b) a grounded base stage is used to give unity current gain and a high output impedance. While the 10Ω video input impedance is too low a load for tuned transformer coupling, the grounded base stage can be run at say 0·25mA by suitable choice of $R_1$, $R_2$, $R_3$ to provide an input impedance of 100Ω and a useful power gain.

Finally, gain control: because the integrated amplifier contains a d.c. loop, the actual emitter currents cannot be changed without upsetting the working conditions. It is however possible to vary the a.c. loop gain as shown in Fig. 16 (c.f. Fig. 9 in Part 1 of this article). The a.g.c. voltage varies the impedance of the diode by varying the

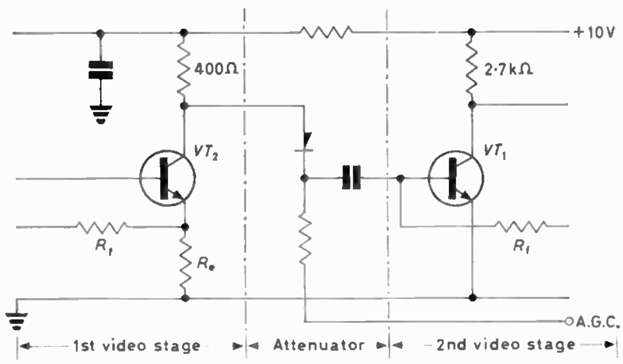


**Fig. 17. A.G.C. applied to feedback pair—series diode attenuator**

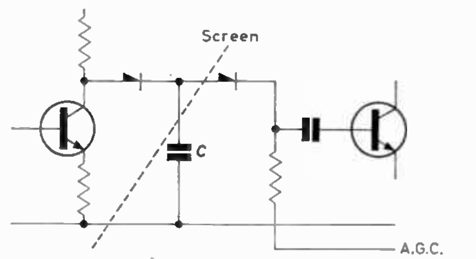


**Fig. 18. Compensation for capacitance of attenuator diode**

current through it. As the diode impedance varies, so does the efficacy of $C$ in by-passing $R$, and hence the gain varies. The d.c. condition is preserved by extending $R_e$ through 1kΩ to a −10V rail (10mA through $VT_2$). The disadvantage of this method is the limited range of control because the stage gain, given at minimum by:

$$\frac{R_f + R_e + R}{R_e + R}$$

can never be made less than unity.

A method giving wider range of control is that using a series diode attenuator as in Fig. 17. Varying the diode impedance by means of its d.c. varies the amount of a.c. fed into the second amplifier. The range of control attainable is 40 to 50dB. A low capacitance diode should be used, and the ZS40 has been found most satisfactory. If the capacitance is appreciable it will shunt the diode resistance when this is biased off and so reduce the control range. Furthermore, over a wide band the capacitance will cause the gain to be higher at the higher frequencies. The effect can be reduced by shunting the 400Ω resistor in Fig. 17 down to 50Ω or less. This enables a given attenuation to be obtained with a lower diode impedance, but causes some loss in maximum gain. Alternatively the circuit of Fig. 18 can be used. Two diodes halve the total diode capacitance, a screen reduces stray capacitance, and the capacitance $C$ found experimentally in the range 10 to 50pF compensates for diode capacitance when biased off, but has little effect when the diodes are low impedance.

**Conclusions**

What has been described is an approach to tuned amplifier design using video amplifiers for the larger part of the overall gain. In particular, using the high frequency integrated amplifier ZLA10 there is the additional advantages of higher frequency performance than can be achieved with conventional components, smaller size, simpler circuits, and a circuit configuration that aids stability and internal feedback. The same circuit element can be used at all frequencies up to 100Mc/s, and with any amount of gain control with no effect on the overall band shape. The number of tuned circuits depends only on the band shape, and is not complicated by having to use additional tuned stages in order to achieve a given overall gain. There are obvious advantages where two or more amplifiers are required having the same band shapes, but different gains or gain control ranges. Conversely, ganged gain controls could be arranged on amplifiers of different gains, band shapes or even centre frequencies. One imagines that the ultimate in design will be crystal filters and (standardized) video amplifiers with the virtual elemination of tuned circuits for all except essential variable frequency applications.

**Future Development of Video Circuit**

The component values as quoted in Fig. 12 are the design values. The actual values obtained in the earlier diffused resistor (integrated) amplifiers were slightly different, resulting in a gain of $\times 7$ (17dB) rather than $\times 10$ (20dB)—hence the quoted 68dB gain for four cascaded stages. The gain of this amplifier depends on the ratio of two resistors: this is necessary when using resistors made by a diffusion process because, while the individual values cannot be reproduced with better than 10 per cent tolerance, the ratio of resistors in any one element is better than 2 per cent.

The peak in the frequency response is due to capacitance across resistor $R_e$, mostly the capacitance under the evaporated aluminium end connexion at the junction of $R_f$ and $R_e$. When evaluation of the initial circuit is complete, new diffusion masks are to be prepared with which this capacitance will be reduced to about a third of its present value. If necessary compensating 'stray' capacitance may be built in across $R_f$. There is one effect that may be difficult to eliminate other than by making separate amplifier blocks. This is the fact that the high frequency gain is dependent to some extent on the collector-base capacitance of the second transistor. This is the Miller effect capacitance which varies with the a.c. gain of $VT_2$ and hence with the effective collector load resistance. As it stands, the optimum value of capacitance that has to be added across $R_f$ to obtain a level response has to be changed slightly depending on whether a high or low load is connected to the output.

Further additions that are being considered for the final design are:

(1) The building in of a decoupling resistor.

(2) Bringing out separately the collector of $VT_2$ and the lower end of the 400Ω resistor to allow choice of damping when used with a tuned circuit.

(3) Building in a larger value tapped resistor in place of $R_f$ to allow the circuit to be used either as a low gain (20dB) high frequency or as a high gain (30dB) lower frequency amplifier.

The noise figure of the present amplifier at 40Mc/s with a 400Ω source impedance is 8dB. At 15Mc/s, 7dB. The transistors used in the earlier amplifier are type ZT2475/2N2475. The final version will use a lower noise transistor similar to the 2N918 at present still being evaluated. The use of this device should reduce the noise figure to about 5dB.

Finally. while the present amplifier is constructed in three parts separately mounted on the header, viz. two transistors and the diffused resistor silicon wafer, it is hoped that eventually it will be possible to combine these on a single wafer. This will involve diffusion isolation of the three regions prior to formation of the transistors and resistors. At the high frequencies involved there will be several major difficulties such as inter-diffusion capacitances and series bulk resistance of the silicon. When these difficulties are solved a major step will have been made towards the complete integration of multi-stage amplifiers. Wolff[15] gives an interesting review of present and future trends towards overall microminiaturization.

**Acknowledgments**

Particular thanks must be extended to Mr. R. Naylor for most of the high frequency theoretical predictions, and the earlier experimental results that led to the development of this amplifier in integrated form.

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## Digital Flight Information Logging

The Sperry Gyroscope Co. Ltd has now completed the development of an entirely new airborne solid state digital data acquisition system, capable of handling from six to many thousands of parameters.

Immediate applications are as an incident/crash recorder; as a flight test recorder; or as a maintenance recorder, for finding faults and indicating trends. Industrial and marine applications are also probable.

The airborne system, briefly described below, has been designed as a series of modules, a combination of which will meet impending air safety regulations and the customer's own requirements for additional parameters.

The timing of all operations in the system is controlled from a crystal-controlled multivibrator, which drives a divider chain of 20 divide by 2 circuits; the states of this chain control the sequences of all operations.

Each input, of whatever form, is converted in a 'pulse-duration convertor (p.d.c.) to the common form of a pulse train where the pulse length is proportional to the magnitude of the inputs; these are multiplexed in a solid state unit whose programme may easily be changed.

A high-frequency square wave from the divider chain is gated during a single pulse from a p.d.c. into a binary counter. The number in the counter is then proportional to the magnitude of the input. Where the information is recorded on magnetic tape, 'non-return to zero' parallel recording is used. The information in the counter is transferred to a register after each count, and the states of the 10 stages of the register are continuously recorded on 10 tracks. A parity bit is recorded on an 11th track, a word synchronization signal on a 12th, and frame synchronization on a 13th.

When the information is recorded on wire, 'non-return to zero' serial recording is used. The first word of each frame is a frame-synchronization signal, and subsequent words (each of 16 bits) comprise a word synchronization signal followed by the recording of the parameter obtained by scanning the stages of the counter. A cross check is obtained by following this with a parity bit.

# A New Technique for the Recognition of Resistance in the Presence of Reactance

By E. R. Wigan, B.Sc., A.M.I.E.E.

*In this article a bridge type technique is described which enables resistance to be recognized and, if required, measured in the presence of reactance. The theoretical basis of the measurement is given in detail and a typical circuit illustrated.*

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

A 'BRIDGE-TYPE' technique for recognizing the resistive element, alone, of a 'lossy' reactance has been devised[1].

Conventional methods involve the measurements of reactance in addition to the resistance, and in consequence demand a frequency-stable signal-source. With both these complications eliminated the new technique becomes simple to apply and to use; the presence of resistance is recognized immediately the test-object is connected; measurement, if required, may take a few seconds.

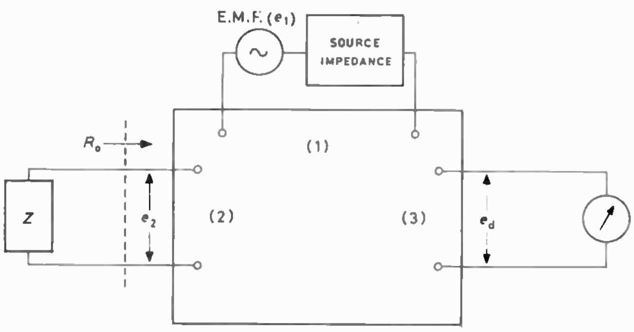


**Fig. 1. The basic arrangement**

*The three-port resistive net is so proportional that (given $e_d = 0$ when $Z = R_b$) $R_b$ is identical with $R_o$—the output impedance of the net measured at port (2)*

Because the network involved has the form of a 'bridge' (see Fig. 1) the new technique can properly be so described; the usage of the bridge, however, is completely unconventional. For example, although the voltage $e_d$ appearing at the 'detector' terminals becomes some measure of the loss-resistance $r$ present in the unknown impedance $Z$, that voltage is never reduced to zero—as in conventional practice—but remains very nearly constant. Measurements of $r$ are made in terms of the *departure from constancy of* $e_d$, or—speaking more strictly—of the ratio $e_d/e_1$, ($e_1$ being the supply e.m.f.).

When the (purely resistive) bridge network is properly proportioned, a short-circuit, or an open-circuit of the test-terminals (2) of the bridge, or the connexion of a completely loss-less reactance, all lead to the same $e_d/e_1$ ratio—phase-relation being ignored.

By adopting one of the many available artifices the constant magnitude of this ratio can be converted into a 'null' reading of the detector. This null is disturbed when the test-object contains resistance $r$: $r$ will be treated as the 'equivalent series resistance'.

The meter reading then becomes nearly proportional to $r$, unless the $r/X$ ratio is substantial. On the other hand, when $X$, alone, is present the deflexion remains at zero irrespective of the sign of the reactance. Such properties strongly recommend the device for mass-production testing, the tested item being rejected if the meter current exceeds a preset maximum.

The circuit conditions set out in the legend of Fig. 1 give all the essential information upon which an '*R*-Recognizing' bridge can be designed. Note, however, that the basic circuit elements are required to be purely resistive. If an '*X*-Recognizing' bridge were required the same theory—*mutatis mutandis*—will apply, the basic elements being purely reactive.

The Appendix deals with the basic circuit theory of the *R*-Recognizer but the practical embodiments, of which there exist an unlimited number, are not dealt with; equally

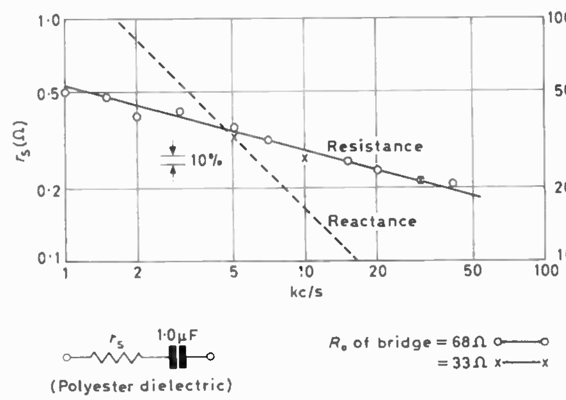


**Fig. 2. Measurements made using polyester dielectric capacitor**

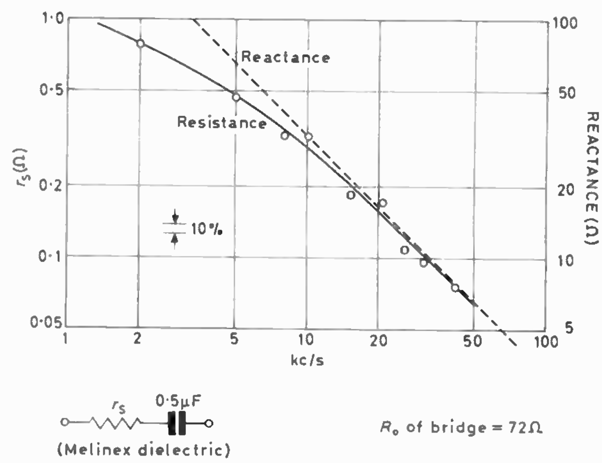


**Fig. 3. Measurements made using Melinex dielectric capacitor**

the various artifices for generating a 'null' detector reading from a constant value of $e_d/e_1$ are left to the ingenuity of the designer.

For various technical reasons the 'ratio arms' of the resistive bridge are best replaced by a tapped transformer-winding. The residual imperfections of the transformer can then be swamped by using resistive attenuating networks to ensure that the essentially resistive elements of the network operate in an effectively resistive milieu. A practical circuit uses a centre-tapped transformer associated with five auxiliary resistors—if the transformer was perfect, three resistors would suffice.

Such an assembly, capable of operating from audio to radio frequencies was demonstrated by Messrs. Hatfield Instruments Ltd at the recent Exhibition of the Physical Society. Some graphs illustrating measurements made with an earlier, low-frequency, model are shown in Figs. 2

and 3. The following circuit information applies to this model.

The idealized network of Fig. 4 illustrates the proportions of the bridge used for tests with $R_o = 33{\cdot}3\Omega$. Here the two idealized sources $E_1$ have zero impedance and the instrument observing $E_3$ takes no current; elements $A$, $C$ and $D$ are pure resistances and $T$ is the impedance of the test-object.

Then the conditions called for (see legend of Fig. 1) are met if: $A = K.C$ and $D = C.(K-2)(K+1)/(K+2)$, which leads to $R_o = C.(K-2)(K+1)/2K$. (Notice that $K > 2$, or $D$ and $R_o$ become negative).

Then when $T = 0$, $E_3 = +E_1/(1+K)$

when $T = \infty$, $E_3 = -E_1/(1+K)$

while when $T = \pm jX$, $|E_3|$ is unchanged.

(if $K = 3$, $|E_3| = E_1/4$.)

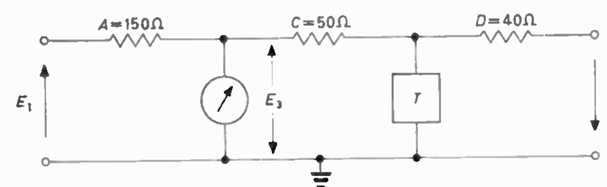


**Fig. 4. Ideal network**

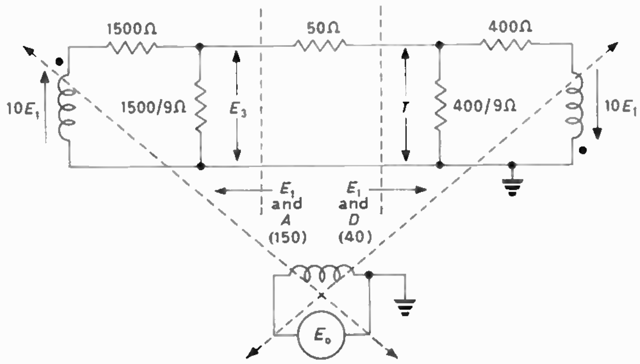


**Fig. 5. Practical network**

In practice the e.m.f's $E_1$ are generated by a transformer (Fig. 5); at the same time attenuating pads (in the figure 10/1 voltage ratio) are used as explained above to 'hold off', from the $A$-$C$-$D$ net, any unwanted impedance contributed by the transformer.

This arrangement has the great practical advantage that a measuring 'meter' of quite low input impedance (provided it is resistive) can be used; the resistor across which it is shunted is increased to restore the value of $A$.

To observe that $|E_3|$ remains constant when $T = 0$ or $\infty$, and is therefore insensitive to anything except resistance in $T$, $E_3$ is amplified, rectified, and a d.c. voltage $V_1$ stored on a capacitor; meanwhile a second alternating voltage (equivalent to $E_1/(1+K)$) is taken from the source $E_o$ and similarly stored as $V_2$. The difference between $V_1$ and $V_2$ operates the final indicating meter in the test-equipment; this meter stands at zero unless $T$ contains resistance.

## Appendix

(1) Consider the influence of $Z$ upon $e_2$ and $e_d$ under the conditions listed below:

| Condition | $Z$ | $(e_2)$ | $(e_d)$ |
|---|---|---|---|
| (a) | $\infty$ | $E_1$ | $D_1$ |
| (b) | $R_b$ | $E_2$ | zero |
| (c) | zero | $E_3$ | $D_3$ |
| (d) | $(r+jX)$ | $E_4$ | $D_4$ |

(2) Then, because $R_b = R_o$,

$$E_2 = (E_1)/2$$

while clearly $E_3$ = zero.

(3) The argument turns on the *changes* in $e_d$ due to changes of $Z$. The change in $e_d$, $\Delta_1$, due to altering $Z$ from $R_b$ to $\infty$ is:

$$\Delta_1 = E_1 - E_2 = +E_1/2$$

Again the change in $e_d$ due to short-circuiting $R_b$ is:

$$\Delta_2 = E_2 - E_1 = -E_1/2$$

(4) The corresponding changes in the 'detector' p.d. $e_d$ must—by the principle of superposition—inevitably reflect (in some fixed proportion) the p.d. changes $\Delta_1$ and $\Delta_2$ which have occurred at terminals (2).

The changes were all relative to the condition that $Z = R_b$, and under that condition (b) $e_d$ = zero.

It immediately follows that:

$$D_1 = -(D_3)$$

(5) It is then easily deduced that when $Z = (r + jX)$, (Condition (d)):

$$D_4/D_1 = M, \text{(say)} = (1 - (r+jX)/R_o)/(1+(r+jX)/R_o)$$

(see footnote) ............ (1)

or:

$$M = (1 - \alpha - j\beta)/(1 + \alpha + j\beta) \quad \text{.......... (2)}$$

where $\alpha = r/R_o$ and $\beta = X/R_o$.

Thus if $\alpha$ (i.e. $r$) is absent $|M|$, the magnitude of the ratio $M$, is unity and is unaffected by $X$. On the other hand $|M| = 1$ again if $\alpha = \infty$.

If $\alpha$ is present, but $\beta$ vanishingly small:

$$|M| = 1 - 2\alpha \quad \text{................ (3)}$$

Or if, as is most common, $\alpha$ and $\beta$ are both finite:

$$|M| \simeq 1 - 2\alpha/(1 + \beta^2) \quad \text{............ (4)}$$

From these equations it is to be seen that $|M|$ will never *exceed* unity and will *fall* from this value only if $\alpha$ (that is $r$) is finite.

By the artifices mentioned this fall in voltage can be converted into a meter deflexion, $N$, which stays at zero unless $r$ is finite.

Thus in the general case of equation (4), $N$ is proportional to $2\alpha/(1+\beta^2)$. Because of the term in $\beta^2$, the reading $N$ is *reduced* by the reactance $X$; nevertheless $N$ remains *proportional* to $\alpha$ and $r$, if $\beta$ (or $X$) is fixed. In consequence the ohmic value of $r$ can be deduced by inserting—in series with the unknown—a known resistance $R_k$ which reduces $N$ to $N/2$. $R_k$ then equals $r$.

The elegance of this substitution technique is that the effect of $\beta$ (or $X$) is completely eliminated except insofar as it reduces the 'sensitivity' of the measurement—see equation (4).

A slightly more sophisticated, direct-reading version of this measurement technique was used in collecting the data plotted in Figs. 2 and 3. The 'spread' of the data around the mean smooth curve shows up the loss of sensitivity as $X$ or $R_o$ were varied. (Two separate bridge networks were used; two resistors only have to be changed to alter $R_o$).

**REFERENCE**


1. Patent Application 43704/63.


*This formula indicates that, when $r = 0$, the network has 'all pass' transmission characteristics. A note on the design and application of such networks is in preparation.*

# A Transistor Amplifier with D.C. and A.C. Feedback Stabilization

By H. C. Bertoya*, A.M.Brit.I.R.E.

*This article describes and analyses a transistor amplifier which uses a d.c. feedback loop in order to stabilize the d.c. working point. If the amount of a.c. feedback required does not coincide with the amount of d.c. feedback it is possible to have two separate feedback loops.*

*Direct coupling is used between stages and this leads to a simple and economical design. Practical examples are given using both pnp and npn transistors. One example was specifically designed for low noise application.*

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

A USEFUL and elegant transistor amplifier has been described by M. K. McPhun[1] which employs a method for stabilizing the d.c. bias current through the first stage. The amplifier is directly coupled, but the way in which the transistors are used does away with the need for any special circuits to allow for different levels of d.c. potential through the amplifier. This leads to a circuit of good performance which at the same time is extremely simple and employs the minimum number of components.

No full analysis of the mode of operation of this circuit was given, so the analysis was carried out and it was found to lead to some useful results.

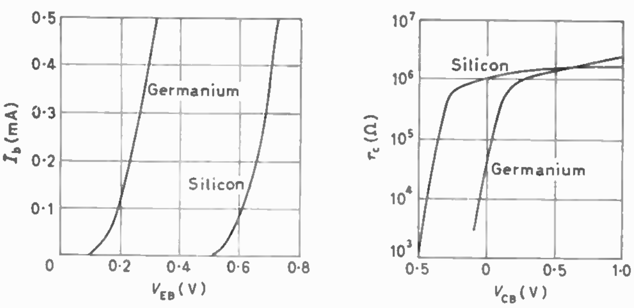


**Fig. 1 (a).** Typical characteristics of emitter-base diodes for alloy junction germanium and silicon transistors

**(b).** Collector resistance as a function of collector-base voltage for typical alloy junction germanium and silicon transistors

### Notes on Symbols

In the analysis which follows $E_{ab}$ or $E_{AB}$ refers to a voltage existing between points a and b or A and B. A and B denote d.c. values and a and b denote a.c. values. b and B are more positive than a and A. Current flows from the negative to the positive end of an impedance. For simplicity $h_{fe}$ is written as $\beta$ and $h_{FE}$ as $B$.

## Mode of Operation

### Silicon Transistor Characteristics

Typical characteristics of a silicon transistor are shown in Figs. 1(a) and 1(b). The curve of Fig. 1(b) shows that the transistor will operate with a small reverse base-collector voltage. Advantage can be taken of this fact in the circuit of Fig. 2 where it is seen that transistor $VT_2$ is directly coupled to $VT_1$ without the need for any intermediate potential.

Although $R_{L1}$ is passing a current $i_{c1}$ as well as $i_{b2}$, point C assumes the potential given by the curve of Fig. 1(a). This will not differ substantially from the potential at B since the curve is steep. If the change of current through $VT_1$ and $VT_2$ does not alter the voltage at C and D beyond the linear operating region, then linear current amplification will be obtained. The matter is discussed fully in the paper by M. K. McPhun[1].

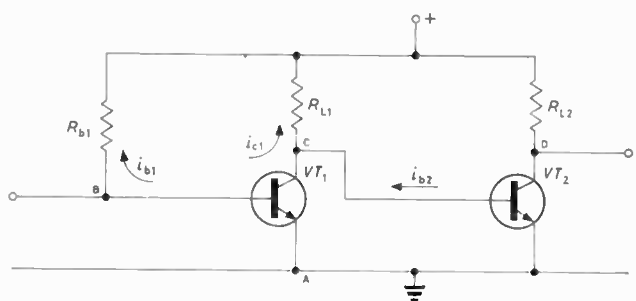


**Fig. 2.** D.C. coupled transistors

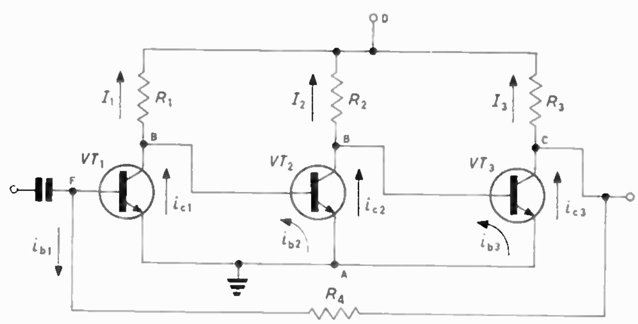


**Fig. 3.** Circuit of basic amplifier

### The Complete Amplifier and its Analysis

The complete basic amplifier is shown in Fig. 3.

To set up the circuit equations for the d.c. conditions it should be noted that the voltage $E_{AB}$ will remain sensibly constant no matter what the value of $i_{b1}$ or $i_{b2}$ (see Fig. 1(a)).

The circuit equations are:

$$i_{c1} = B_1 i_{b1},\quad i_{c2} = B_2 i_{b2},\quad i_{c3} = B_3 i_{b3},$$
$$i_{b2} = I_1 - i_{c1},\quad i_{b3} = I_2 - i_{c2}.$$
$$I_1 = (E_{AD} - E_{AB})/R_1,\quad I_2 = (E_{AD} - E_{AB})/R_2,$$
$$I_3 = i_{c3} + i_{b1}$$
$$E_{AC} = E_{AD} + E_{DC},\quad E_{DC} = -R_3(i_{c3} + i_{b1})$$

Thus:

$$E_{AC} = E_{AD} - R_3(i_{c3} + i_{b1}) \quad \ldots\ldots\ldots (1)$$

By substituting successively for $i_{c3}$ and $i_{c2}$:

$$E_{AC} = E_{AD}[1 - R_3B_3/R_2 + R_3B_3B_2/R_1] + E_{AB}[R_3B_3/R_2 - R_3B_3B_2/R_1] - i_{b1}[R_3 + R_3B_1B_2B_3] \ldots (2)$$

---

* *British Scientific Instrument Research Association.*

But:

$$i_{b1} = (E_{AC} - E_{AF})/R_4 \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

Substituting equation (3) in equation (2) and re-arranging yields:

$$E_{AC} = \frac{E_{AD}R_4\left[R_1R_2 + R_3B_3(R_2B_2 - R_1)\right] + E_{AB}R_3R_4B_3(R_1 - R_2B_2) + E_{AF}\left[R_1R_2R_3(1 + B_1B_2B_3)\right]}{R_1R_2\left[R_4 + R_3(1 + B_1B_2B_3)\right]} \quad \ldots\ldots \quad (4)$$

This is the complete equation. $E_{AF} \simeq E_{AB}$ but its identity has been maintained in the equation so that its effect may be more readily apparent.

In all practical cases one may assume that $B_1B_2B_3 \gg 1$, and the equation may be rewritten accordingly. If it is assumed, in addition, that $R_3B_1B_2B_3 \gg R_4$ and $R_2B_2 \gg R_1$ then the equation may be written as:

$$E_{AC} = E_{AD}R_4\left[1/R_3B_1B_2B_3 + 1/R_1B_1\right] - E_{AB}R_4/R_1B_1 + E_{AF} \quad \ldots\ldots\ldots\ldots \quad (5)$$

Now suppose $R_3B_2B_3 \gg R_1$, then:

$$E_{AC} = (E_{AD} - E_{AB})\,R_4/R_1B_1 + E_{AF} \quad \ldots\ldots \quad (6)$$

Thus, given $R_1$ and $B_1$, it is possible to obtain the required voltage at C by selecting a suitable value of $R_4$.

If a satisfactory circuit is to be built whose performance accords with equation (6), it is necessary to examine the four approximations which follow equation (4), and the way in which the circuit must be proportioned in order to make the approximations valid. These points will be taken in order.

(a) $$B_1B_2B_3 \gg 1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

As previously noted, relation (7) is easily satisfied.

(b) $$R_3B_1B_2B_3 \gg R_4 \quad \ldots\ldots\ldots\ldots\ldots \quad (8)$$

The minimum value of $R_3$ is given by relation (8) but if a low value of amplifier output resistance is required then $R_3$ must be low. $R_4$ is generally of a high value so that if a very low output impedance is required an emitter-follower stage may be necessary.

(c) $$R_2B_2 \gg R_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

This implies that $R_1$ and $R_2$ must be components with the same order of resistance. It is not possible to 'grade' the sequence of resistor loads as is usually done when the first stage passes a very small current.

(d) $$R_3B_2B_3 \gg R_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (10)$$

This may be readily achieved.

A reasonable sequence of values for $R_1$, $R_2$ and $R_3$ would be in the ratios 10, 10 and 1.

### Variations due to Temperature

The value of $I_{co}'$ in silicon transistors is small and if the standing current through the first transistor is made large in comparison then the effect of temperature may be considered by noting the change of $E_{AC}$ with change of $B_1$.

Consider an open loop d.c. coupled amplifier with $R_4$ returned to a stable voltage $E_{AG}$.

The performance may be calculated from equation (2). The current $i_{b1}$ is now given by:

$$i_{b1} = (E_{AG} - E_{AF})/R_4 \quad \ldots\ldots\ldots\ldots \quad (11)$$

Substituting equation (11) in equation (2) gives:

$$E_{AC} = E_{AD}\left[1 + R_3B_3/R_2 + R_3B_3B_2/R_1\right] + E_{AB}\left[R_3B_3/R_2 - R_3B_2B_3/R_1\right] - E_{AG}\left[(R_3 + R_3B_1B_2B_3)/R_4\right] + E_{AF}\left[(R_3 + R_3B_1B_2B_3)/R_4\right] \quad \ldots\ldots\ldots\ldots \quad (12)$$

$$\therefore\ dE_{AC}/dB_1 = -(E_{AG} - E_{AF})\,R_3B_2B_3/R_4 \quad \ldots\ldots \quad (13)$$

Considering the closed loop amplifier given by equation (6) then:

$$dE_{AC}/dB_1 = -(E_{AD} - E_{AB})\,R_4/R_1B_1^2 \quad \ldots\ldots\ldots \quad (14)$$

It will be seen that the quantity $dE_{AC}/dB_1$ is very much reduced in the circuit configuration described by equation (14).

$B_1$ is a function of temperature so the variation of $E_{AC}$ may be computed by suitable substitutions in equation (14). Typical curves for a Mullard BCZ11 and a Texas Instruments 2N929 are given in Fig. 4.

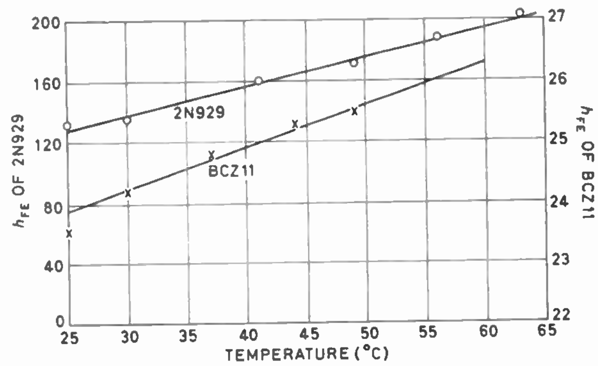


Fig. 4. Variation of $h_{FE}$ with temperature

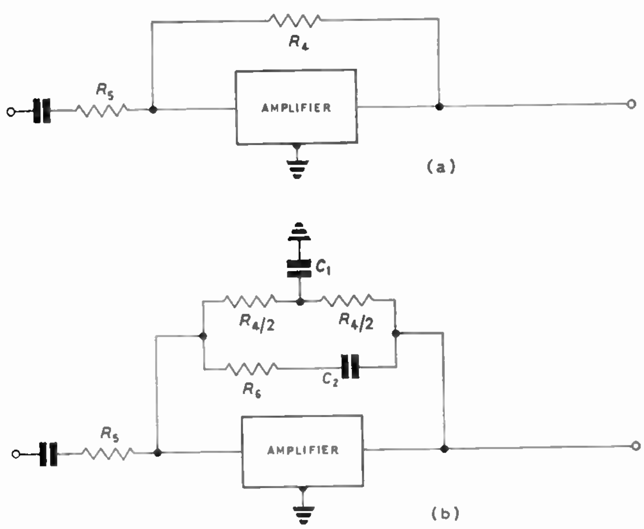


Fig. 5 (a). Amplifier with common d.c. and a.c. feedback
(b). Amplifier with separate d.c. and a.c. feedback

### D.C. and A.C. Feedback

The circuit of Fig. 3 shows the amplifier with the d.c. feedback path through $R_4$. Fig. 5(a) shows the block diagram of a complete amplifier. The d.c. feedback path consists again of $R_4$. The amount of a.c. feedback depends approximately on the ratio of $R_4$ to $R_5$.

If it is not possible to proportion the circuit so that $R_4$ is satisfactory for both the d.c. and a.c. feedback then separate networks may be used as in Fig. 5(b). The components $R_4$, $C_1$ comprise the d.c. feedback and $R_5$, $R_6$ and $C_2$ the a.c. feedback.

### A Practical Example (Amplifier No. 1)

An example of an amplifier designed according to the preceding theory is shown in Fig. 6. The additional components $C_4$, $C_6$ were inserted to prevent loop instability.

### CALCULATIONS OF $E_{AC}$

The required component values are:

$R_1$ and $R_2 = 10\text{k}\Omega$, $R_3 = 1\text{k}\Omega$, $R_4 = 120\text{k}\Omega$

$B_1 = 24$, $B_2 = 35$, $B_3 = 43$

$E_{AF} = -0{\cdot}6\text{V}$, $E_{AD} = -12{\cdot}25\text{V}$, $E_{AB} = -0{\cdot}64\text{V}$

(the amplifier uses pnp transistors)

Inserting these values in equation (6) gives $E_{AC} = -6{\cdot}4\text{V}$. The measured value was $-6{\cdot}36\text{V}$. The parameters were all measured at the same temperature.

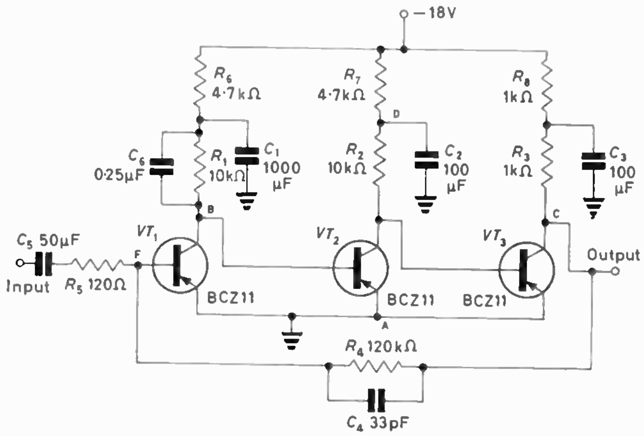


**Fig. 6. Circuit diagram of amplifier No. 1**

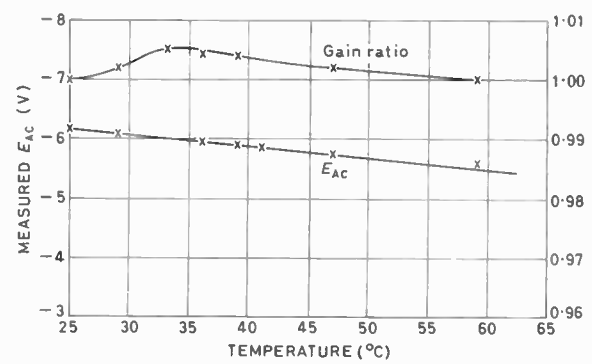


**Fig. 7. Variation in performance with temperature (amplifier No. 1)**

### CALCULATIONS OF $dE_{AC}$

For a temperature range of 22°C to 68°C an approximate mean value of $B_1$ taken from Fig. 4 is 24·8. The value of $dB_1$ is 2·4. By inserting values in equation (14) $dE_{AC} = +0{\cdot}54\text{V}$ is obtained. This compares with a measured value of 0·64V. The inaccuracies arise from the change with temperature of the values of $E_{AB}$, $E_{AF}$, $R_1$ and $R_4$ which have not been considered. Nevertheless, the order of variation is readily obtainable. It is not, of course, employed as a design parameter.

### PERFORMANCE DATA FOR AMPLIFIER No. 1 (FIG. 6)

| | |
|---|---|
| Gain | 850 |
| Gain (open loop) | 23 000 |
| Output resistance | 25Ω |
| Frequency response (to −3dB) | 40c/s to 30kc/s |
| Stability margin | 12dB |

Variation of performance with temperature is shown in Fig. 7.

The a.c. gain is plotted as the ratio:

Gain at any temperature/gain at lowest temperature.

All a.c. measurements were carried out at 1kc/s and an output voltage of 1V r.m.s. The source impedance was 10Ω.

## A Practical Example (Amplifier No. 2)

This amplifier has separate d.c. and a.c. feedback loops as described previously; otherwise it is similar in design to amplifier No. 1. The circuit is shown in Fig. 8. The components $C_6$, $C_7$ were for the purpose of loop stabilization.

### PERFORMANCE DATA FOR AMPLIFIER No. 2

| | |
|---|---|
| Gain | 3 300 |
| Gain (open loop) | 40 000 |
| Output resistance | 80Ω |
| Frequency response (to −3dB) | 40c/s to 27kc/s |
| Stability margin | 12dB |

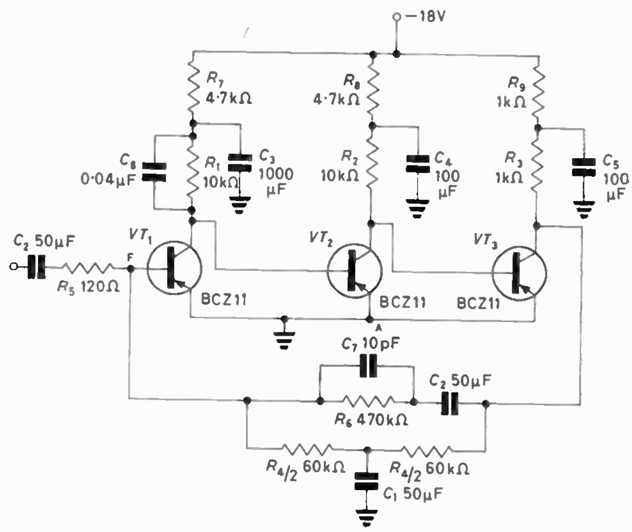


**Fig. 8. Circuit diagram of amplifier No. 2**

Variation of performance with temperature is shown in Fig. 9.

All a.c. measurements were carried out at 1kc/s and an output voltage of 1V r.m.s. The source impedance was 10Ω.

## A Practical Example (Amplifier No. 3)

This unit was required to operate as a high-gain, low-noise pre-amplifier with a very low output impedance. The circuit of Fig. 6 was modified to include an emitter-follower. If the analysis is carried out to include this extra stage it is found that the d.c. potential between emitter and base occurs on both sides of the equation, so long as the output voltage is measured at the emitter. There is thus no need to modify the result of equation (6).

### CALCULATION OF $E_{AC}$

The required component values are:

$R_1$ and $R_2 = 100\text{k}\Omega$, $R_3 = 10\text{k}\Omega$, $R_4 = 5\text{M}\Omega$

$B_2 = 34$, $B_3 = 44$

Three 2N929 transistors were connected in the input stage

**TABLE 1**

| $E_{AF}$ | $E_{AB}$ | $E_{AD}$ | $B_1$ | CALCULATED VALUE OF $E_{AC}$ (V) | MEASURED VALUE OF $E_{AC}$ (V) |
|---|---|---|---|---|---|
| (a) 0·58 | 0·59 | 12·7 | 128 | 5·28 | 5·4 |
| (b) 0·57 | 0·59 | 12·5 | 116 | 5·7 | 5·8 |
| (c) 0·57 | 0·59 | 12·75 | 95 | 6·97 | 7·08 |

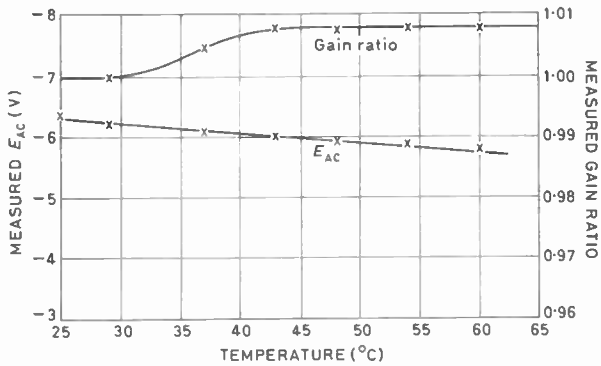


**Fig. 9. Variation in performance with temperature (amplifier No. 2)**

each transistor and this is shown in Fig. 11. All a.c. measurements were carried out at 1kc/s and an output of 1V r.m.s. The source impedance was 10Ω.

The capacitor $C_6$ (0·01μF) and $C_7$ (5pF) are for loop stabilization. A further two amplifiers were built and their open loop gain was found to be higher—about 100 000. In this latter case the capacitors were changed to 0·025μF for $C_6$ and 22pF for $C_7$. The bandwidth was about 35kc/s and the stability margin was greater than 20dB.

The performance of a typical amplifier at low temperatures is shown in Table 2.

### Amplifier No. 3 Driven from a High Impedance Source

This amplifier was driven from a source with 1MΩ out-

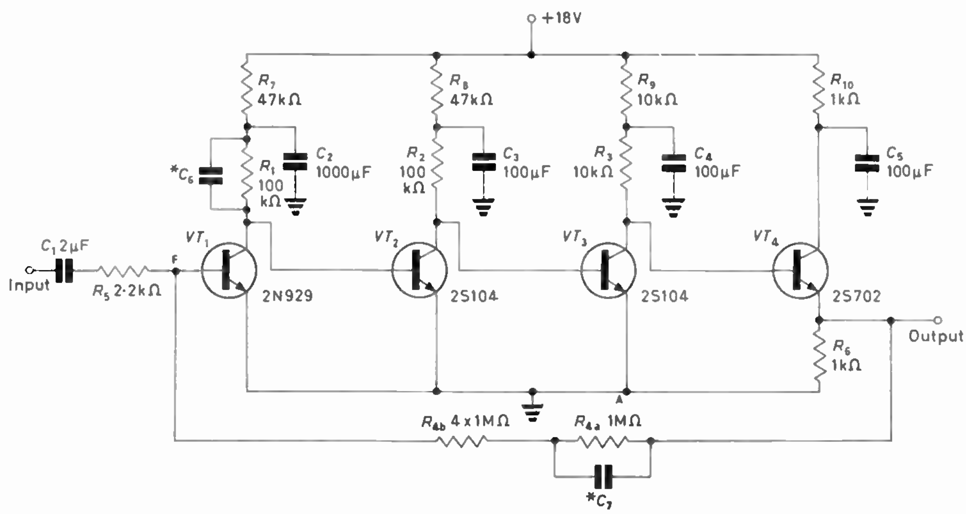


**Fig. 10. Circuit diagram of amplifier No. 3**

The various values of $B_1$, measured and calculated values of $E_{AC}$, etc., are given in Table 1. In this case the transistors are npn.

#### Calculation of $dE_{AC}$

From case (b) for a temperature range of 24°C to 56°C $B_1$ changes from 118 to 188. Inserting the relevant values into equation (14) gives $dE_{AC} = -1{\cdot}77$V, which compares with a measured value of $-1{\cdot}4$V.

#### Performance Data for Amplifier No. 3 (Fig. 10)

(The same performance was obtained with all three transistors.

| | |
|---|---|
| Gain | 2 200 |
| Gain (open loop) | approx. 50 000 |
| Output resistance | approx. 9Ω |
| Frequency response | 35c/s to 30kc/s |
| Stability margin | > 20dB |

Noise with source impedance short-circuited 4·5mV r.m.s. approx.

The performance with temperature varied slightly with

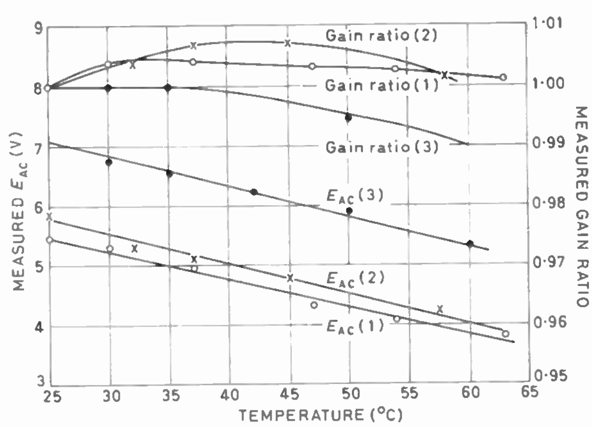


**Fig. 11. Variation of performance with temperature (amplifier No. 3)**

put impedance. The amplifier was terminated with a 1kΩ load.

The amplifier oscillated when driven from the 1MΩ source. The oscillation was stopped by the alteration of $C_6$ to a value of 0·1μF.

#### Performance Data for Amplifier No. 3 Driven from High Impedance Source

| | |
|---|---|
| Gain ($E_{ac}/i_s$) | = 5 × $10^6$V/A |
| Bandwidth | = below 10c/s to 40kc/s |
| Stability margin | = 10dB |

Performance with respect to temperature is given in Fig. 13.

**TABLE 2**

| TEMPERATURE (°C) | GAIN RATIO | $E_{AC}$ (V) |
|---|---|---|
| +31 | 1 | 3·95 |
| +22 | 1 | 4·3 |
| −1 | 1 | 4·9 |
| −7 | 1·002 | 5·2 |
| −12 | 1·002 | 5·4 |
| −15 | 0·998 | 5·6 |

Fig. 13 was plotted with the 1MΩ source resistor at a constant temperature. When the resistor was placed in the oven along with the amplifier the change in gain ratio was from unity at 25°C to 1·009 at 85°C.

## Miscellaneous Notes

### Value of $B$ ($h_{FE}$)

$B$ ($h_{FE}$) varies considerably with the value of the collector voltage when this is very low. The measurement of $h_{FE}$ must therefore be carefully made. Typical curves for the BCZ11 and 2N929 transistors are given in Fig. 13.

### Input Capacitor

The choice of electrolytic or paper capacitors depends on the value of $R_4$. If good quality electrolytic capacitors are used the maximum value of $R_4$ can be around 470kΩ. Fifteen used and unused Hunt type L37/1 25V 100μF capacitors were placed in an oven and their equivalent resistance was measured with 9V applied. This was always greater than 45MΩ at 20°C and 20MΩ at 50°C.

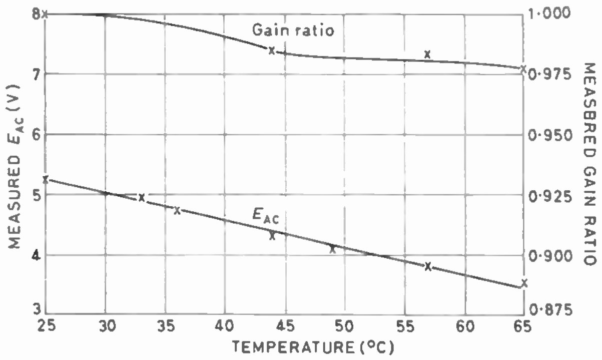


Fig. 12. Variation of performance with temperature (amplifier No. 3 driven from 1MΩ)

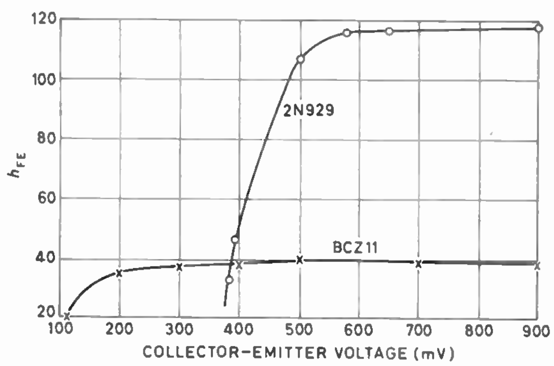


Fig. 13. Variation of $h_{FE}$ with collector voltage for BCZ11 and 2N929

### Calculation Accuracy

The accuracy with which a performance can be predicted depends, of course, on the accuracy with which the various circuit constants are known. In the case of amplifier No. 3 the high values of $B$ ($h_{FE}$) and $R_4$ make accurate calculation difficult.

### Variation of $E_{ac}$ (a.c. Voltage Output)

$E_{ac}$ should increase continuously with temperature since $h_{fe}$ increases. The performance of the amplifiers is better than might be expected because of the negative temperature coefficient of the feedback resistor. This was particularly apparent in amplifier No. 3, where originally $R_{4a}$ and $R_{4b}$ were a single 5MΩ resistor, not of a high stability type. The gain ratio fell continuously with temperature from unity at 22°C to 0·94 at 60°C. This resistor was replaced by 5 × 1MΩ high stability resistors and the gain varied in the manner shown in Fig. 12. The variation of $E_{AC}$ was also affected.

### Low Frequency Stability of System with a.c. and d.c. Stabilization

Suppose the d.c. feedback loop has been decoupled in order to allow separate a.c. feedback and that the source impedance $R_S$ is coupled into the amplifier via a capacitor. The circuit can be drawn as in Fig. 14. If $R_S$ tends to zero it will be seen that there are two $RC$ networks in the loop and the system will have a damped oscillatory response when disturbed. When a.c. feedback is applied, however, point F becomes a virtual earth and the effect of the $RC$ network at the input becomes negligible. If only a small amount of a.c. feedback is envisaged it is advisable to employ common a.c. and d.c. feedback and dispense with $C_1$.

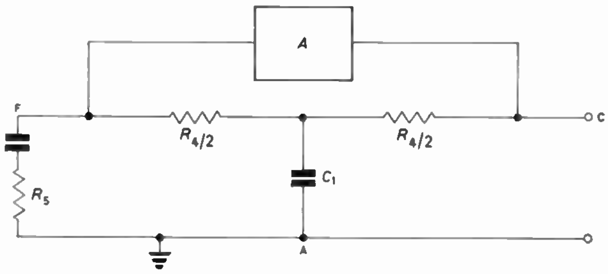


Fig. 14. Basic amplifier circuit with the d.c. feedback decoupled

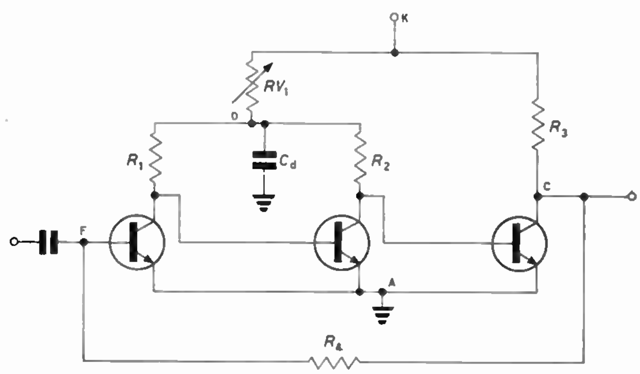


Fig. 15. Method of adjusting $E_{AC}$

### Compensation for Differing Values of $h_{FE}$

It may be inconvenient to have separate a.c. and d.c. feedback loops in very high gain amplifiers. (Amplifier No. 3, for example uses five 1MΩ resistors in the feedback loop). Suppose a fixed value of $R_4$ is required in order to keep the a.c. gain constant from amplifier to amplifier. It is possible for a particular value of first stage $h_{FE}$ to cause the d.c. output voltage to be either high or low so that, with large signals, distortion can occur at the output stage. If this is the case, the d.c. output voltage may be adjusted by alteration of $R_1$. $R_1$ must not differ too greatly from the values given in the analysis and equation (9) must always be satisfied.

### Feedback and Circuit Stability

Since the circuit has no series coupling capacitors it is possible to employ large amounts of feedback with only simple stabilizing networks. For example, one amplifier had an open loop gain of about 900 and a closed loop gain of unity. The bandwidth was 30kc/s, and the stability margin greater than 12dB. The stabilizing network consisted solely of a capacitor across the first collector load.

**Conclusion**

The design procedure for amplifiers with a.c. and d.c. feedback has been described. The procedure may be summarized as follows:

(1) Determine the value of $h_{FE}$ of the first transistor.

(2) In the case of voltage amplifiers, proportion $R_4$ and $R_5$ together to give the correct value of d.c. output voltage and a.c. gain. In the case of current amplifiers it is only necessary to proportion $R_4$. Reasonable accuracy may be achieved by putting $E_{AF}$ and $E_{AB}$ equal to 0·6V in equation (6). If the a.c. output voltage is small and $E_{AD}$ is large then considerable latitude is tolerable in the value of $E_{AC}$.

(3) If the requirements for a.c. gain and the value of $E_{AC}$ are incompatible, use separate a.c. and d.c. feedback loops.

(4) In practice it has been found that feedback amplifiers with substantially identical characteristics may be constructed simply by completing the circuit except for $R_1$. $R_1$ is then chosen (a 1 per cent high stability resistor but from the 10 per cent range) to give the closest approximation to the desired value of $E_{AC}$.

(5) Yet another method of adjusting $E_{AO}$ is seen in Fig. 15. Analysis of the circuit yields the same result as equation (6) where the d.c. output is independent of $E_{AK}$. The desired value of $E_{AC}$ is obtained by adjustment of $RV_1$ (which alters the value of $E_{AD}$). The components $RV_1$ and $C_1$ together are also used to form a decoupling network.

**Acknowledgments**

The author wishes to acknowledge the valuable assistance rendered by Mr. A. R. Ramsay who built the amplifiers, measured their performance and checked the calculations. Acknowledgment is also made to the Director and Council of the British Scientific Instrument Research Association for permission to publish this article.

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# An Improved Technique for Measuring Skin Resistance

By L. E. H. Lindahl*, A. H. Christianson* and R. N. Gooch*

*A method of measuring skin resistance using a substitution technique is described. The balanced d.c. amplifier employed overcomes the non-linearity, lack of sensitivity and inadequate output of many existing instruments. In addition, it provides automatic calibration. Experience has shown that its stability is well within the tolerance limits required in psychophysiological work.*

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

APPARATUS limitations may be responsible for errors and inadequacies in studies using psychogalvanometers. An examination of previous apparatus shows that the main weaknesses have been:

(1) Non-linearity.

(2) Lack of sensitivity.

(3) Low output.

The apparatus to be described here, completely overcomes these limitations and also provides additional advantages of great value.

The salient features characterizing this improved psychogalvanometer are, briefly, as listed below.

(1) The linearity of this apparatus is always the same for any value of resistance. The actual departure from linearity depends on the base level helical potentiometer used and the manufacturers estimate this to be $\pm 0{\cdot}1$ per cent. This potentiometer permits discrimination, in this case to 250Ω, i.e., one thousandth part of the total range.

(2) Its design ensures that range switching if required brings in no spurious effects.

(3) Continuous calibration accurate to $\pm 1$ per cent is inherent in the design and technique regardless of the absolute value of the subject's resistance.

(4) The provision of a range of sensitivity settings gives a maximum sensitivity one hundred times greater than that given by the psychogalvanometers referred to above. It is possible to assign objectively accurate measurement to deflexions as low as 250Ω but deflexions of 50Ω or less are objectively detectable, (deflexions of less than 50Ω are clearly observable but less reliable).

(5) The characteristics of the output of this instrument, coupled with its overall sensitivity, lead to the important advantage that the standard pen recorder is rendered capable of following changes in resistance much more accurately than the previous instruments and moreover changes resulting in deflexions of small amplitudes, such as indicated above, which occur in some psychogalvanic reflexes, may be readily recorded.

**Description of Apparatus**

Fig. 1, the circuit diagram, shows the electronic details. Four double triodes are used in a direct coupled amplifier, the input stage of which is a balanced cathode-follower arrangement with a stepped voltage divider for selection of appropriate sensitivity ranges. The figures on the right-hand side of the divider (1kΩ, 5kΩ, etc.) show the choice of sensitivities available. The next two stages are a balanced cascade voltage amplifier which feeds a balanced cathode-follower power output stage. The instrument is

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essentially a valve-voltmeter bridge, and it measures the voltage difference between the two input grids.

The power supply is a simple neon stabilizer providing 150V and 300V stabilized and 400V unstabilized. $V_1$ and $V_2$ heaters are supplied from a 6·3V winding centre-tapped to earth while $V_3$ and $V_4$ are supplied from a winding centre-tapped to +150V.

## Method of Operation

The apparatus should be switched on about thirty minutes before calibration is attempted so that any drifting due to temperature increases has ceased. The sensitivity control should be in the 'off' position, the electrodes 'shorted', (input switch at position 3) and the base dial should read zero resistance (i.e., the 250kΩ Helipot at maximum). By means of the balance dial the meter needle is set to read zero. Should the needle not remain at zero the apparatus must be allowed further time to warm up and attain complete stability.

When this is achieved the sensitivity control is switched to the 5kΩ position and the 'set zero' control knob is used to bring the meter needle back to the zero position if it has moved during this operation.

After setting zero the sensitivity control is now switched to the 10kΩ position, say, and the base dial rotated to read 10kΩ. The meter, which should now show full scale deflexion, may be adjusted to indicate this value, if not attained, by means of the 'full scale' control knob. The instrument is now calibrated.

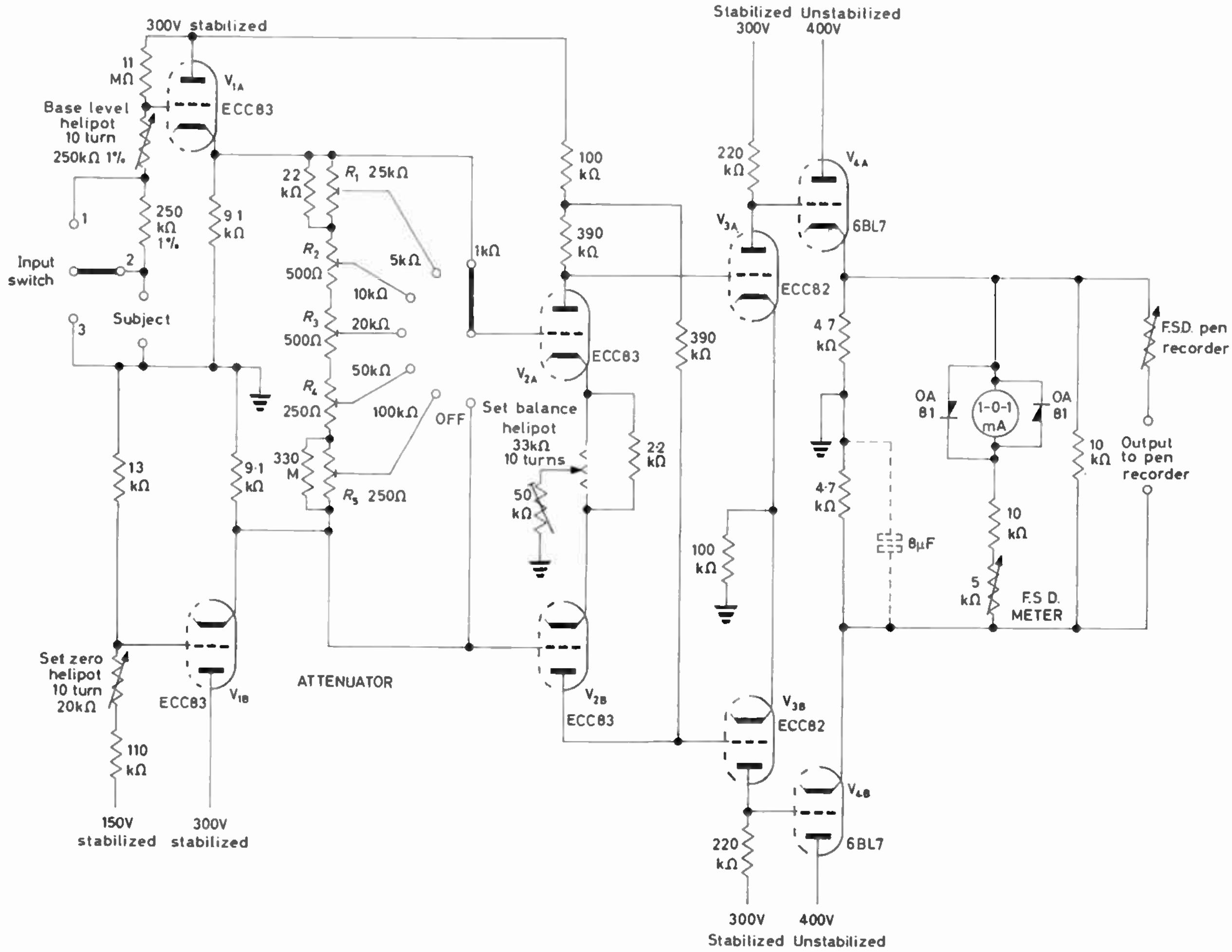


**Fig. 1. The complete circuit**

The recorder may be aligned with the meter by means of the recorder adjustment knob.

After setting up in the manner described, the subject's resistance is introduced in series with the resistance chain by turning the input switch to position 2. The value of the subject's resistance is now obtained by rotating the dial of the Helipot until the meter reads zero. The reading on the Helipot dial indicates how much resistance has been taken out of the resistance chain (now including the subject) necessary to restore balance. Changes in the subject's resistance from this basal level are indicated by the meter and/or the pen recorder.

The attenuator range switch selects the degree of sensitivity at which it is desired to indicate these changes in subject resistance. Position 1 of the input switch is used with subjects having a skin resistance greater than 250kΩ. The maximum range for basal resistance measurement is thus extended up to 500kΩ. The constant current through the subject is 26$\mu$A.

The method of attaching the electrodes to the skin found to give the best results is the one suggested by Malmo.[1]


**Acknowledgment**

Grateful acknowledgments are due to Dr. C. M. Franks for supplying the basic circuit from which this apparatus was developed.


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# Silicon Avalanche Rectifiers

By G. Duddridge*

*The manufacture and the characteristics of silicon avalanche rectifiers are described and it is pointed out that this type of rectifier has the ability to absorb reverse over-voltage transients. Both series and parallel operation are dealt with and applications are described.*

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

SILICON power rectifiers are now an accepted part of many power supply systems which derive d.c. from a.c. Their compactness, efficiency and low cost are well known, and their ability to operate at high temperatures has enabled considerable progress to be made in the reliability of d.c. power supplies.

However, all conventional silicon rectifiers with the possible exception of silicon reference diodes (Zener diodes) have required protection against any voltage transient likely to exceed the peak transient voltage rating of the device. Over-voltage protection has normally consisted of a capacitive-resistive surge absorption network, or the use of specially processed selenium plates. Both these solutions have increased the basic cost of rectifier equipment, have required expensive and intensive testing and have introduced further problems, for example the difficulty of voltage doubling when using *CR* networks.

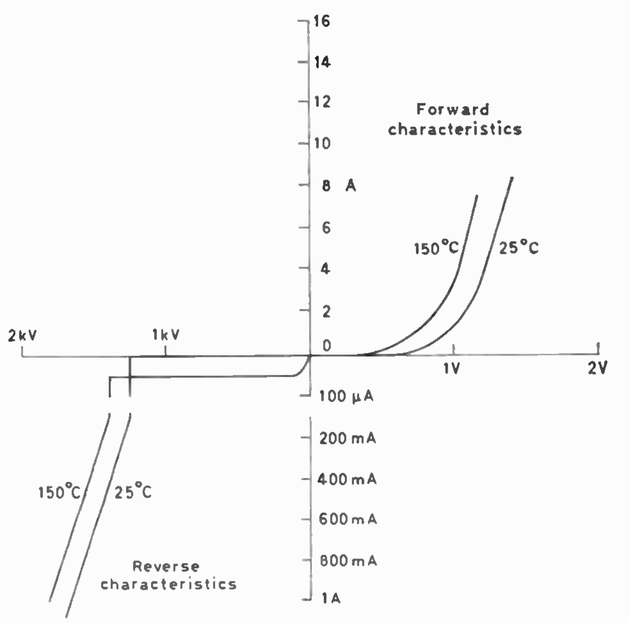


**Fig. 1. Avalanche rectifier characteristics**

The present day approach to this problem is the introduction of the silicon avalanche rectifier. This diffused junction silicon rectifier is especially processed to ensure that the reverse voltage breakdown occurs in the bulk material of the junction and not at one small point on the surface. Such a mechanism gives the rectifier the ability to absorb reverse over-voltage transients.

### Manufacture of Silicon Avalanche Rectifiers

The main problems of current lead connexions to diffused wafers were overcome several years ago and many all-diffused devices, including Thyristors (silicon controlled rectifiers) are now available. The special surfaces required for silicon avalanche rectifiers presented a considerable technical production problem; for what seems easy to the research laboratory can be very difficult when production of quantities at relatively low cost is envisaged. However, the special problems of these devices have now been solved, and the RAS310 avalanche rectifier is the first product of a complete new line of silicon avalanche power rectifiers. This rectifier is manufactured in only one voltage grade, which is calculated to meet the needs of most light current applications. It has a forward current rating of 1·25A with normal air cooling.

### Characteristics of Controlled Avalanche Rectifiers

In the forward direction, these silicon avalanche rectifiers exhibit properties identical to conventional silicon rectifiers, having low forward losses and operating at high current densities. In the reverse direction the new rectifiers have a clearly defined avalanche breakdown voltage characteristic, with very low leakage current at high temperatures, and behave as a high voltage Zener diode (Fig. 1). Avalanche rectifiers have a dynamic slope resistance, which is a function of the surface area. Typically for the RAS310 this is 500Ω, or approximately $5 \times 10^{-4}\,\Omega^{-1}/\text{mm}^2$ of junction area. The rectifiers have a positive temperature coefficient of avalanche voltage of approximately 0·1 per cent/°C. Probably the single most important characteristic of silicon avalanche rectifiers is the transient thermal resistance (Fig. 2). This, coupled with two junction temperature ratings (the maximum operating junction temperature of 140°C and the maximum transient junction temperature of 220°C) and the forward and reverse characteristics, defines completely the rectifier rating. Work in this field is at an early stage, but a tentative rating of the RAS310 using the method is given later.

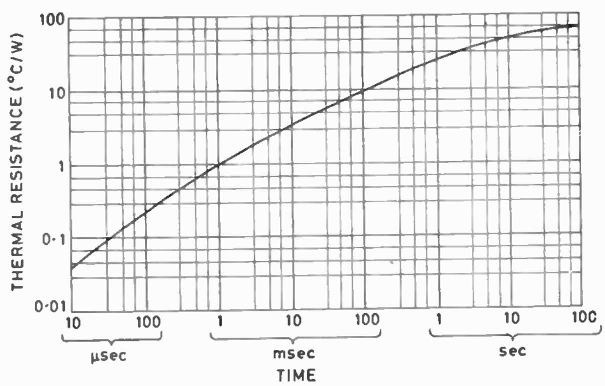


**Fig. 2. Transient thermal resistance**

### The Rating of Silicon Avalanche Rectifiers

The normal forward current rating of avalanche rectifiers is determined by the steady state thermal resistance, forward characteristics, and peak operating junction temperature[1]. For maximum forward current it is necessary to ensure that no appreciable reverse losses occur. Hence in

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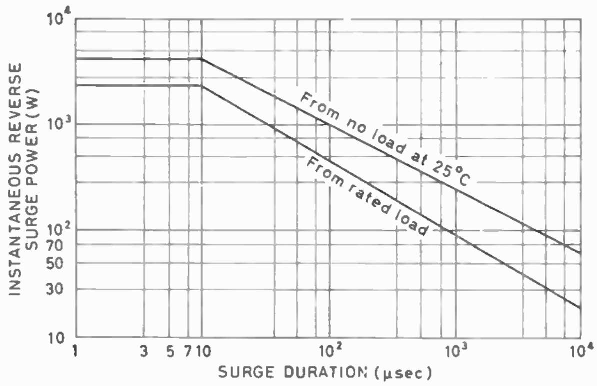


**Fig. 3. Surge reverse power**

normal rectifier circuits the rated crest of the r.m.s. operating voltage is set 20 per cent below the avalanche voltage, and the peak recurrent voltage is set just 4 per cent below the minimum avalanche voltage at 25°C. Any high voltage transients which occur will cause reverse power losses which will increase the rectifier junction temperature, but provided these are not too frequent, and are below the on-load reverse surge rating (Fig. 3), they will be absorbed by the rectifier.

When a device is not used at its maximum current rating the new surge rating can be readily calculated, as the following example shows.

**Example**

For a rectifier whose characteristics are:

dissipation at 1A direct forward current = 1W
thermal resistance = 75°C/W
transient thermal resistance at 100μsec = 0·25°C/W
reverse avalanche voltage = 1 000V
reverse slope resistance = 500Ω

determine the maximum reverse transient square wave power which the device can withstand in the reverse direction for 100μsec at an ambient temperature of 45°C where the rated maximum transient junction temperature is 220°C. Determine also the current and voltages present.

Operating junction temperature = (ambient temperature) + (power loss × thermal resistance)
= 45 + 1 × 75
= 120°C

Permissible transient rise = 220 − 120
= 100°C

∴ transient power dissipation = (100/0·25)W
= 400W

Thus $400\text{W} = I_R V_A + I_R^2 R_A = 1000\,I + 500\,I^2$

from which $I \simeq 340\text{mA}$
$V \simeq 1\,175\text{V}$

i.e., the rectifier in this application will reduce a 400W transient of 100μsec duration to approximately 1 175V without damage. This method of rating, by using the superposition theory can also be applied to trains of pulses[2].

If silicon avalanche rectifiers are to be used in the avalanche mode of operation continuously, the limits defined by the thermal resistance curve, and the peak operating junction temperature must be very carefully observed. The peak transient junction temperature cannot be used as a continuous rating without degradation of the device.

It must be pointed out that for temperatures below 25°C the avalanche voltage will decrease, to approach the peak repetitive voltage rating.

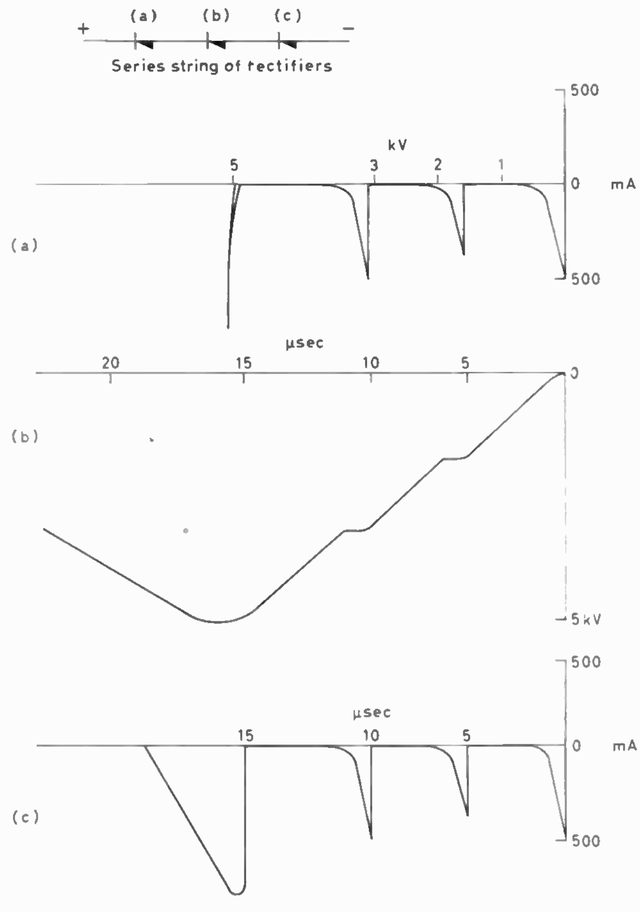
**Series Operation**

The basic factor to be considered in the problem of series operation is the power dissipation which takes place within the device having the fastest recovery time of those in the series chain.

For low power operation (10 per cent of rated current), series strings of silicon avalanche rectifiers require no resistive or capacitive divider networks when operating at frequencies up to 25kc/s with sinusoidal waveform. For higher powers (i.e. up to 100 per cent rated current) or square waveforms, hole storage equalizing capacitors are at the moment considered necessary. Voltage sharing resistors are completely unnecessary. Fig. 4 shows the curve trace of three silicon controlled avalanche rectifiers in series, being switched from forward into reverse conduction, and the voltage/time and current/time waveforms. The $V/I$ trace can be explained as follows. Rectifier (b) has the lowest recovery time and recovers first. The voltage builds up across (b) until this reaches its avalanche voltage, (b) avalanches and (a) has not yet completely recovered. Current flows through (b) until (a) reverts to a blocking state. The voltage now builds up across (a) and (b) until it is sufficient to drive (a) and (b) into avalanche, thus completing the recovery of (c) and the whole chain reverts to a blocking state. This sequence of events can be avoided by selecting rectifiers for identical recovery time characteristics—and eliminating hole storage equalizing capacitors—but this is not at the moment considered to be an economic proposition.

Once again transient voltage suppression is considered unnecessary because of the inherent ability to operate

**Fig. 4 (a). Voltage/current characteristics**
**(b). Voltage/time characteristics**
**(c). Current/time characteristics**



reliably in the reverse avalanche breakdown region.

High voltage (10kV) stacks or avalanche rectifiers are giving satisfactory service in particular applications without surge suppression of equalizing components. 100kV stacks are currently under consideration for an application in which the capacitance to ground is more important than hole storage equalization.

**Parallel Operation**

The operation of two or more silicon avalanche rectifiers in parallel in the forward direction is exactly the same as for a conventional silicon rectifier. Forward derating, or the use of current sharing reactors or series resistance are acceptable methods for current equalization.

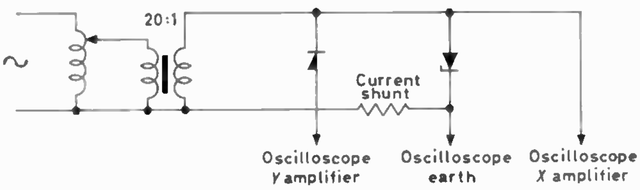


Fig. 5. Forward test circuit

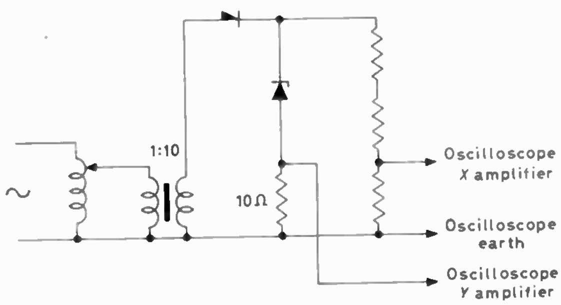


Fig. 6. Reverse test circuit for avalanche voltage

When devices are considered in parallel in the reverse direction, a more complex problem arises. If one device has a characteristic avalanche voltage considerably below that of the remainder, this device may be required to handle all the power from a transient voltage surge, with the possibility of excessive power dissipation and finally destruction. The slope resistance of the avalanche region and the positive temperature coefficient of the avalanche voltage act as a safety factor, but for long duration surges with correspondingly lower avalanche currents, it is essential that the avalanche voltage characteristics do not differ excessively. This requires that the factory must supply specially selected avalanche rectifiers where parallel operation is required. Wherever possible the next higher current rating device should be used without parallel connexion.

Even where the avalanche voltage characteristics differ by only 30V, a 20 per cent derating on permissible avalanche transient power should be applied to parallel devices. Where matching of reverse characteristics is not carried out the permissible reverse surge power is that of one device for up to 10 devices in parallel, and that of two devices for 10 to 20 devices in parallel, and so on.

**Applications of Silicon Avalanche Rectifiers**

These diodes are primarily intended for rectification purposes, but when specially selected—owing to their unique high voltage avalanche properties—they can also be used for applications requiring very high voltage Zener diodes. The most important application is the protection of sensitive electronic circuits from high voltage surges, or the preventon of forward break-over in s.c.r. circuits.

Avalanche rectifiers can successfully be used in conjunction with Zener diodes to stabilize e.h.t. voltages at any required voltage level.

Silicon avalanche rectifiers can also be used for fast rise time, high voltage pulse formation, the only limitations being that of permissible power dissipation to maintain peak junction temperatures below 140°C. These rectifiers have been successfully used to produce 1 300V peak-to-peak square waves of 100$\mu$sec duration, with rise times of the order of 0·1$\mu$sec.

Transient voltage surges can be very simply detected and monitored using a number of specially selected silicon avalanche rectifiers with known avalanche voltages.

**Test Circuits for Silicon Avalanche Rectifiers**

The forward characteristics of silicon rectifiers can be measured, using either d.c. or a.c., in any conventional circuit such as that shown in Fig. 5.

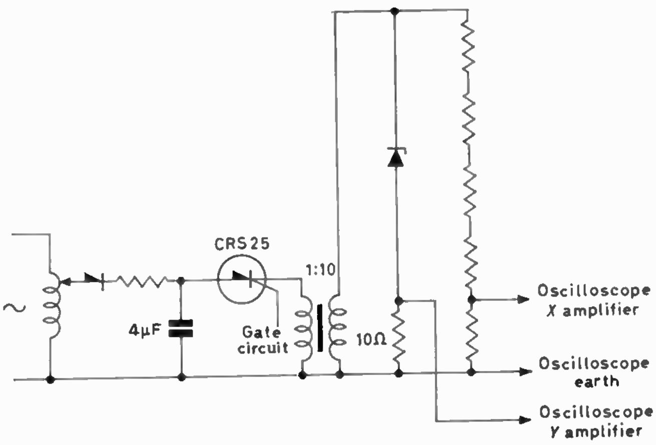


Fig. 7. Reverse test circuit for avalanche characteristics

The reverse avalanche voltage can be measured using the a.c. technique outlined in Fig. 6. D.C. measurements of greater accuracy may be made, but it is essential that the continuous dissipation ratings are not exceeded in these circumstances.

The stability of the avalanche characteristic may be shown by the circuit given in Fig. 7. This measurement may be single-shot or repetitive, giving also the avalanche dynamic resistance. High speed *X* and *Y* amplifiers are essential, together with a frequency compensated potential divider network.

**Conclusions**

Silicon avalanche rectifiers can be expected to replace many conventional silicon rectifiers during the next few years; the extra care in processing and testing is expected to show further improvement in the reliability of silicon rectifier devices. Voltage surge suppression networks can in general be considered obsolete where suitably rated avalanche rectifiers are used with an isolating transformer.

The evaluation of silicon avalanche rectifiers—in particular the reverse surge rating—presents a considerable problem to the manufacturer, but it is anticipated that these problems will be completely overcome shortly and a standardized evaluation method evolved.


**Acknowledgment**

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



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# A Broad-Band Tapered Field Resonance Isolator

By J. H. Collins*, M.Sc., and T. M. Heng*, B.Sc.

*This article describes work undertaken on the design of a broad-band resonance isolator using a transversely varying external d.c. magnetic field. A simple theory for the required magnetic field taper is given. This is compared with the field gradient obtained from the fringing field between the poles of a magnet. A ratio of reverse to forward loss of 80 has been achieved experimentally, over the frequency range 7·5Gc/s to 11·5Gc/s, with a 0·300in by 0·100in by 3·25in F5X magnesium-manganese ferrite slab mounted in standard X-band waveguide.*

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

THE basic mode of operation of the ferrite resonance isolator has been known for some years. When a thin ferrite slab is placed in a rectangular waveguide, non-reciprocal attenuation takes places, providing the applied magnetic field is of the correct magnitude to produce ferromagnetic resonance. By properly locating the ferrite within the waveguide, a large difference in the transmission loss for the two directions of propagation is observed. This difference in transmission loss is due to the fact that for one direction of propagation the microwave magnetic field is circularly polarized in the same sense as the free electron spin precession in the ferrite giving pronounced absorption, while for the other direction of propagation the polarization is in the opposite sense. Thus, at ferromagnetic resonance, there is a large absorption (termed the reverse loss) in one direction, and only a small absorption (termed the forward loss) in the other.

During the period of the development of the resonance isolator, numerous suggestions have been put forward for extending its bandwidth. One particular technique, due to Soohoo[1], depends on biasing the ferrite slab with a transverse spatially varying d.c. magnetic field. This gives rise to a continuum rather than a single resonance frequency in the sample, thereby increasing the bandwidth of the device. It is the object of this article to discuss the theory and experimental results on a ferrite isolator, built by the authors, utilizing this principle.

### Theory of the Tapered-Field Isolator

Consider the configuration as given in Fig. 1, in which a ferrite slab is located in the H-plane of a rectangular guide. Optimum performance, namely the largest value of reverse to forward loss, requires that the mean position of the ferrite be dependent on the shape and size of its cross-section[2]. However, for maximum reverse attenuation at any frequency, the point of circular polarization of the internal r.f. field in the ferrite must coincide with that of the empty waveguide. In some geometries this condition cannot be satisfied but for the very thin ferrite slab, this is theoretically justifiable. The plane of circular polarization is then given by:

$$X = L/\pi \tan^{-1} \{(f/f_o)^2 - 1\}^{-\frac{1}{2}} \quad \ldots\ldots\ldots\ (1)$$

where $f_o$ is the cut-off frequency for the empty waveguide, $L$ the guide width and $X$ is measured from a narrow wall of the waveguide. Over a frequency band, therefore, it is desirable to have the planes of circular polarization occurring between $A$ and $B$ (see Fig. 1), where $A$ and $B$ represent the planes at the upper and lower ends of the frequency spectrum.

* *The University, Glasgow*

The resonance d.c. magnetic field, $H_a$, is determined by Kittel's equation. When combined with equation (1) this gives for a thin slab magnetized in the $y$-direction.

$$H_a = f_o/\gamma \operatorname{cosec} (\pi X/L) + 4\pi M \quad \ldots\ldots\ (2)$$

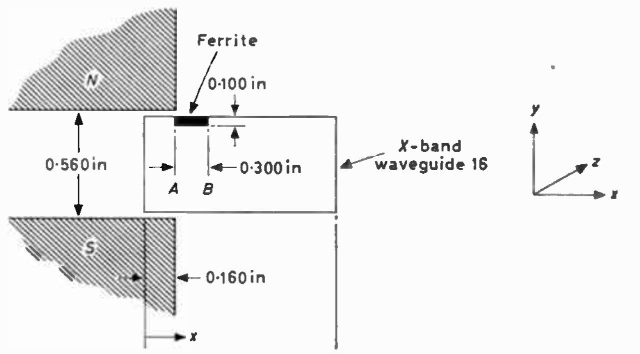



Fig. 1. Broad-band ferrite resonance isolator

where $4\pi M$ and $\gamma$ are the saturation magnetization and the gyromagnetic ratio respectively of the ferrite. Equation (2) gives the theoretical field gradient required to produce the resonance frequency continuum suggested by Soohoo. It is demonstrated in the following section that this gradient can be closely realized using the fringing field of a simple form of magnet.

### Design of Magnetic Field Taper

The solution for the field intensity at any point on the axis of symmetry of the magnet configuration shown in Fig. 2 has been given by Gibbs[3]. By a slight modification the actual magnetic field applied to the ferrite slab, oriented with respect to the magnet as in Fig. 1, can be obtained. Taking from symmetry considerations the $x$-axis as one of the boundaries and using the Schwarz-Christoffel transformation it is found that:

$$dz/dt = g/2\pi\, (t + 1)^{\frac{1}{2}}/t \quad \ldots\ldots\ldots\ (3)$$

which on integrating yields:

$$2\pi z/g = 2(t + 1)^{\frac{1}{2}} - 2 \ln \{(t + 1)^{\frac{1}{2}} + 1\} + \ln t \ldots\ (4)$$

the origin of co-ordinates being at the junction of the symmetry axis and the pole edge plane. The complex potential $W$, appropriate to this problem is given by:

$$W = U + jV = 1/\pi \ln t \quad \ldots\ldots\ (5)$$

where $U$ and $V$ are the stream and potential functions respectively.

Substituting equation (4) into equation (3):

$$2\pi z/g = 2(e^{\pi W} + 1)^{\frac{1}{2}} - 2 \ln \{(e^{\pi W} + 1)^{\frac{1}{2}} + 1\} + \pi W \quad (6)$$

The magnetic intensity, $H'$, is determined by $|dW/dz|$ and thus from equations (3) and (5):

$$H' = 2/g\left|\frac{1}{(t+1)^{\frac{1}{2}}}\right| = 2/g\,|H|(\text{say}) \quad \ldots\ldots \quad (7)$$

Substituting equation (7) into equation (4):

$$2\pi z/g = \left[(2/H) - \ln\frac{1+H}{1-H}\right] \quad \ldots\ldots\ldots\ldots \quad (8)$$

It is convenient to introduce the following definitions:

$$H = R\,e^{j\theta} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

and

$$T\,e^{j\psi} = \left(\frac{1+H}{1-H}\right) \quad \ldots\ldots\ldots\ldots \quad (10)$$

giving:

$$T^2 = \frac{1 + R^2 + 2R\cos\theta}{1 + R^2 - 2R\cos\theta} \quad \ldots\ldots\ldots\ldots \quad (11)$$

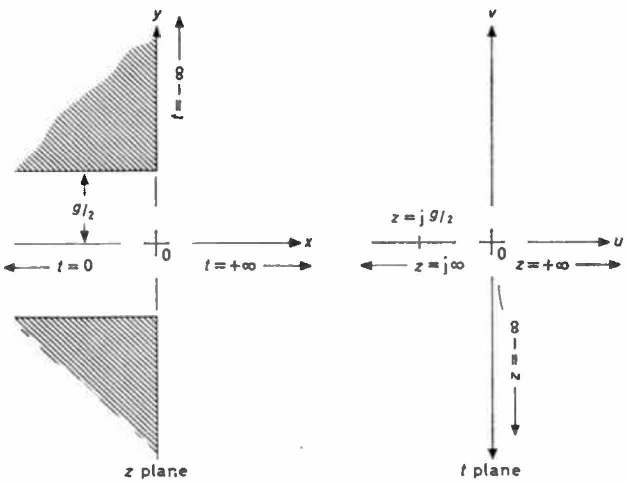



**Fig. 2. Transformation of points in the z-plane to the t-plane**

$$\text{and } \psi = \tan^{-1}\left(\frac{R\sin\theta}{1 + R\cos\theta}\right) + \tan^{-1}\left(\frac{R\sin\theta}{1 - R\cos\theta}\right) \quad \ldots\ldots\ldots\ldots \quad (12)$$

Equation (8) may be rewritten:

$$2\pi z/g = 2/R\,(\cos\theta - j\sin\theta) - \tfrac{1}{2}\ln T^2 - \tfrac{1}{2}j\psi \quad \ldots\ldots \quad (13)$$

For calculation purposes the ferrite, shown in Fig. 1, is considered to be such that its mean displacement from the symmetry axis in the $y$-direction is 0·150in. Further, the effective magnetic field, acting on the ferrite in isolator applications, is $R\cos\theta$. This field is determined as a function of $x$ by equation (13) together with equations (11) and (12). The results for the fringing field, at various values of $X$, are shown in Fig. 3, using a uniform gap field of 7 000 oersted. These are also compared with the theoretical field variation, defined by equation (2) for a continuum in the ferrite resonance. The F5X magnesium-manganese ferrite slab under consideration has the following measured properties:

$4\pi M = 2\,500$ oersted

$\gamma = 2{\cdot}95$Mc/s sec-oersted

$\Delta H = 380$ oersted.

## Experimental Results

Two slabs of ferrite were used in the experiments. Slab A had dimensions of 0·300in by 0·100in by 3·25in and slab B had dimensions 0·100in by 0·050in by 2·50in. Each was located with a narrow edge 0·160in from a narrow wall of the waveguide, see Fig. 1. Unfortunately, premature saturation of the electromagnet used in the experiments prevented the required gap field intensity (7 000 oersted) being obtained. Nevertheless, with a field of 4 600 oersted, reverse attenuations in excess of 20dB were obtained for slab A over the frequency range 7·5 to 11·5Gc/s. The forward attenuation was less than 2dB and substantially of the order of 1dB over 2·7Gc/s of the band investigated. These results are shown in Fig. 4. The match of the system was remarkably good, considering that no special precautions were taken, being less than 1·07 over the defined frequency range.

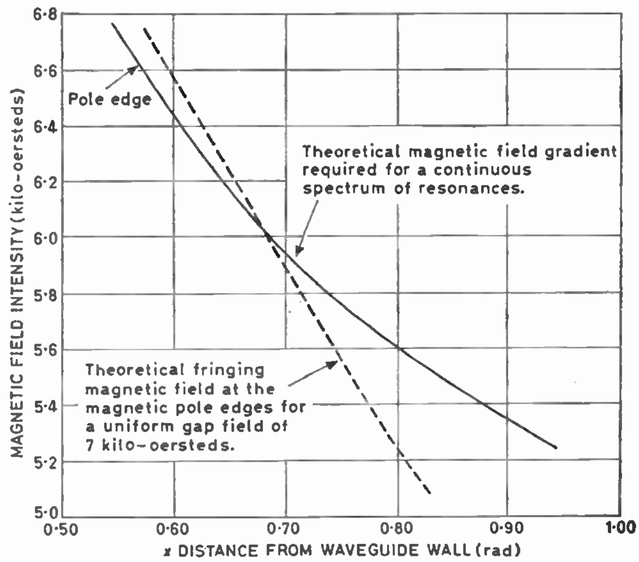
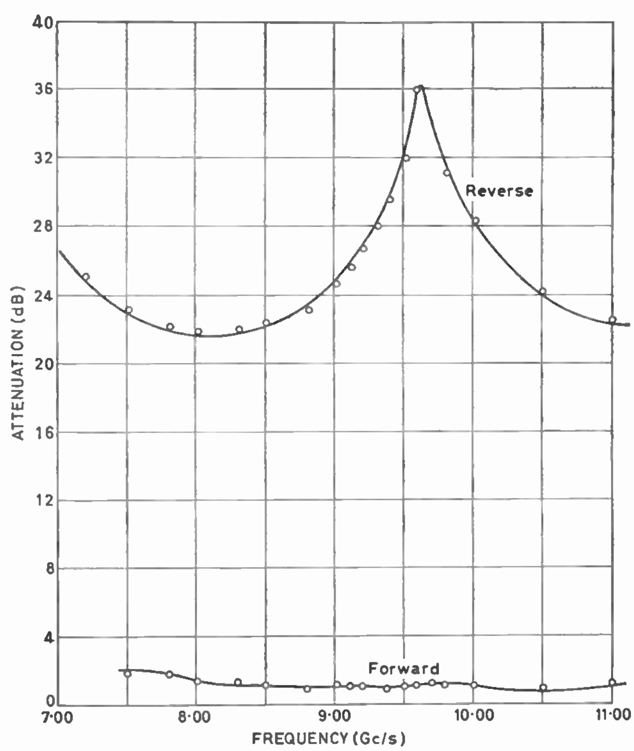



**Fig. 3 (above). Comparison of required magnetic field gradient to the fringing field of a magnet for an F5X ferrite slab of $4\pi M_s$ : 2500Oe, $\gamma$ : 2·95Mc/s/Oe**

**Fig. 4 (below). Measured forward and reverse loss of a 0·300in × 0·100in × fringing field of a magnet for an F5X ferrite slab of $4\pi M_s$ : 2500Oe, 4600Oe**



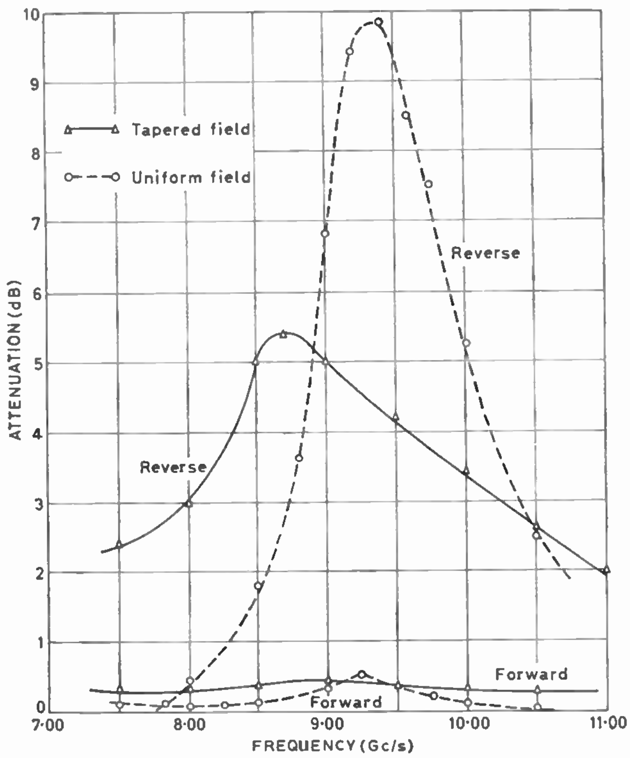



**Fig. 5. Measured forward and reverse loss characteristics of a 0·100in × 0·050in × 2·50in F5X ferrite slab with a uniform and tapered d.c. magnetic $H_a$ : 4600Oe**

Performance figures were taken for slab B using both uniform and tapered fields and these are shown in Fig. 5. It is observed that substantial reduction of forward loss, using a tapered field, occurs compared with slab A for the price of greatly reduced reverse loss. This arrangement, unless made prohibitively long, is therefore not acceptable as a ferrite isolator. The effect of tapering, however, is evident, namely reduced reverse loss coupled with increased bandwidth. Finally, it should be pointed out that, in forming these experiments, little attention has been paid to optimization. The main objective was to verify the feasibility of such a device.

### Conclusions

A ferrite with low saturation magnetization is necessary to keep the required magnetic field to a reasonably low magnitude. It should then be possible to use a permanent magnet whose size and weight does not exceed those in common use on narrow band resonance isolators. Alternatively, the waveguide height can be narrowed, with the additional advantage of field concentration and reduced interaction length, if a wideband transformer to full height guide is designed.

An optimum size of ferrite can be argued from the results obtained through slab A and slab B. For a ferrite of small cross-sectional area, slab B, the low forward loss is obtained at the expense of the reverse loss. In the case of slab A the converse is true. The high forward loss of slab A arises from the dielectric constant of the ferrite giving desirable field concentration but, at the same time, undesirable conductive losses. This can be partially counteracted by replacing some of the ferrite by a strip of a pure dielectric material having a high dielectric constant and low loss. Additionally, high dielectric materials, whether pure dielectrics of ferrites, disturb the pure circular polarization conditions of the empty waveguide, hence increasing the forward loss.

For the taper obtainable from the pole edges used, a material with a linewidth between 100 and 300 oersted is desirable to derive a reasonable reverse loss. Referring to Fig. 3, if the linewidth of the ferrite is greater than the maximum deviation between the two curves, sufficient loss can be obtained without having to bias the ferrite to ferromagnetic resonance. This large value of $\Delta H$, however, lowers the figure of merit, $\omega/\Delta H$, of the device. Thus, if the tapered field requirements can be satisfied, a low linewidth material is desirable.

**Acknowledgments**

The authors wish to acknowledge the facilities made available for carrying out this work by Professor J. Lamb in the Electrical Engineering Department, University of Glasgow, and to Messrs. Ferranti Ltd, Edinburgh, for supplying the ferrite material.

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## A New Low-Cost E.D.P. System

The National Cash Register Co Ltd has announced the addition to their range of a new, low-cost electronic computer —the NCR 449.

Costing under £16 000 for the basic system, the solid-state, stored-program 449 is capable of all the calculation and decision-making of stored-program data processing systems.

Externally, the 449 is similar in appearance to NCR's electro-mechanical accounting machines. But the pedestals on either side of the console contain control circuits, logic boards, and a ferrite core internal main memory of 200 words, with a capacity of 12 decimal digits per word.

The NCR 449 retains the simplicity of operation and the immediately-accessible printed records of the accounting machine. But by employing the internal memory, stored-program concept of the computer, the 449 achieves extended decision-making powers, and has the flexibility to carry out a variety of different operations without the necessity of changing internal mechanisms or logic structure.

Based on the larger NCR 390, with which it is fully compatible and to which it can be readily expanded, the NCR 449 shares the 390's ability to provide immediately accessible 'hard-copy' records that can be read by people as well as by the computer. These records take the form of ledger cards measuring 20in × 16in, bearing conventional printed figures and also holding up to 216 digits of information on magnetic stripes bonded to the back of each card.

Data and instructions can be fed to the 449 through the console keyboard and via the magnetic ledger cards—which also constitute an external memory capable of indefinite expansion. This facility provides the internal memory with greater effective capacity because all variable data is held in the magnetic cards and only introduced into the processor as needed—on a random-access basis.

As a means of computer input, the magnetic cards afford extreme flexibility of programming, enabling the internally-stored program to be overlaid by reading-in subsequent program steps from the cards, providing unrestricted program length, and enabling modification to be made through the magnetic card as well as the console.

The 449's combination of program-controlled operation and use of dual-language records provides the smaller business with a compact low-cost system that is self-contained and capable of meeting all accounting and statistical needs on the spot, economically and at high speed, while producing conventional written records for consultation as and when required.

The 449 system can have the additional feature of an automatic ledger-card reader which will operate under program control at a speed of 2 000 to 2 500 cards per hour.

The NCR 449 provides four methods of output: on magnetic-stripe ledger cards, punched card, punched paper tape, and visible printed records through the integral printer.

# The Synthesis of Cyclic Code Generators

By P. E. K. Chow*, B.Sc., and A. C. Davies*, B.Sc.

*For digital frequency division and code generation, feedback shift registers have a number of advantages over the more well known techniques using binary counters. Methods of designing feedback shift registers are explained and a new method is proposed. This method leads on to the design of a similar but more general circuit arrangement than the feedback shift register, capable of a wider field of application and not limited to the generation of chain codes.*

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

THE best known methods of digital frequency-division and code generation make use of binary counters, the basic building-block being the bistable flip-flop (Fig. 1). Each stage divides the input frequency by two, so that division by powers of two is readily obtained (Fig. 2).

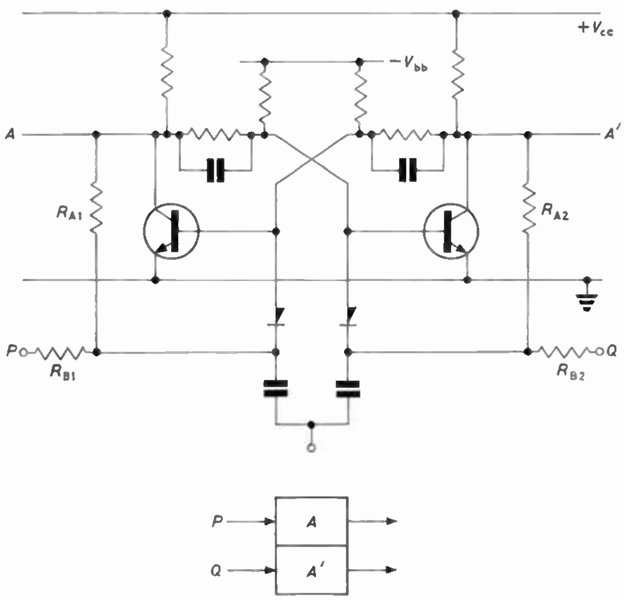


**Fig. 1. A practical circuit and symbolic diagram**

By the addition of feedback loops, this arrangement may be used to divide by other numbers (e.g. ten).

However, such methods suffer from a number of disadvantages. Each stage introduces a delay, so that there is a timing error between the input and output waveforms, which is likely to be significant in equipment operating near the maximum usable frequency of the stages. The versions with feedback loops are prone to spurious transients in their output waveforms. The variety of waveforms that can be generated directly is limited, even using feedback loops.

An alternative arrangement which overcomes these disadvantages is the feedback shift-register (f.s.r.) (Fig. 3).

### Feedback Shift Registers

Fig. 3 shows a five-stage f.s.r. On applying a 'shift' pulse the state of each stage (1 or 0) is transferred one stage to the right, the original state of the last stage ($E$) being lost, and the state of stage $A$ being some Boolean function $S$ of the original states of $A$, $B$, $C$, $D$ and $E$, as determined by the 'logic'. The time of operation is essentially the switching time of a single stage, regardless of the number of stages to the shift-register.

* *The General Electric Co. Ltd.*

The behaviour of the f.s.r. is determined by the 'logic' fed back to the first stage. For example, suppose there are three stages, $A$, $B$ and $C$ and that the 'logic' fed back to $A$ is $S = (A'B' + AC')$. Then, if the initial state is (0, 1, 1) then $S = 0$. Therefore the next state is (0, 0, 1). $S$ is 1, so the next state is (1, 0, 0), and so on. A repeating cycle of five states is produced:

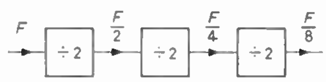


**Fig. 2. Arrangement of the binary counter**

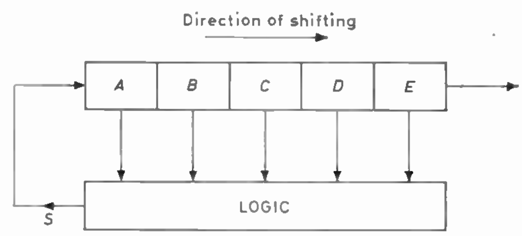


**Fig. 3. A 5-stage feedback shift-register**

| *ABC* | *Decimal Equivalent* |
|---|---|
| 0 1 1 | 3 |
| 0 0 1 | 1 |
| 1 0 0 | 4 |
| 1 1 0 | 6 |
| 1 1 1 | 7 |

If the initial state is (0, 0, 0) or (0, 1, 0), the next state is part of the main cycle. (1, 0, 1) is followed by (0, 1, 0).

The complete behaviour is best visualized by means of a 'state diagram'. For the example chosen, this is shown in Fig. 4. Each circle represents a state of the f.s.r., the number within the circle being the decimal equivalent of that state. The succession of states on a repeating cycle is termed a 'chain code'.

If the feedback function $S$ is entirely a modulo-2 function of the states $A$, $B$, $C$, $D$, $E$ .... then the f.s.r. is termed 'linear', i.e. in this case $S = (C_1A\,(+)\,C_2B\,(+)\,C_3C\,(+)\,C_4D\,(+)\,C_5E\,(+)\ \ldots)$ where $C_1$, $C_2$ .... are either '0' or '1'.

SYMBOLS

| | | |
|---|---|---|
| $A'$ | = | not $A$ = complement of $A$ |
| $A\,.\,B$ | = | $A$ and $B$ together. Logical AND |
| $A + B$ | = | $A$ or $B$ or both. Logical OR |
| $A(+)B$ | = | $A$ or $B$ but not both. Modulo-2 sum of $A$ and $B$ (Equivalent or $AB' + A'B$) |
| $n$ | = | Number of stages in a shift-register |
| $(I)$ | = | unit matrix |

The use of the term 'linear' arises because the mathematical theory of such f.s.r. (in terms of modulo-2 algebra) is closely analogous in form to the algebra of linear networks. An introduction to this theory is given in the Appendix. A more general account is given by Elspas[1].

It will often be found that the state diagram consists of more than one part. Generally, there will be a long 'major' cycle and one or more 'minor' cycles. For example, consider a five-stage linear f.s.r. with $S = D (+) E$ (Fig. 5). The state diagram is shown in Fig. 6. Although f.s.r.'s having state diagrams of this form are of mathematical interest, they are of little practical value, since operation in the major cycle cannot be guaranteed. The cycle of operation will depend on the initial state the f.s.r. takes up when switched on. If such an f.s.r. is to be used, additional logic is essential in order to suppress the unwanted cycles. Practical f.s.r. designs must be such as to ensure self-starting operation in the required cycle.

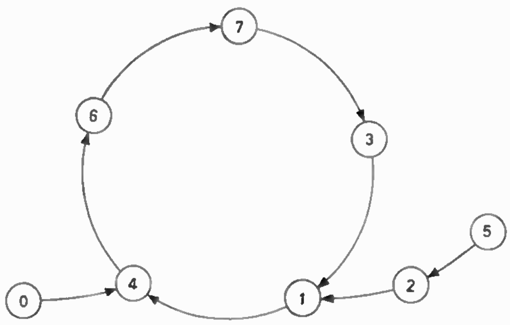


**Fig. 4. A state diagram for S = A'B' +AC'**

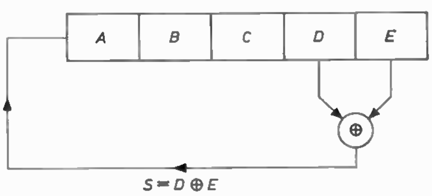


**Fig. 5. A 5-stage linear feedback shift-register with S = D (+) E**

Certain f.s.r.'s will not give cyclic behaviour at all. For example, for a five-stage f.s.r. with $S = (C + E)$, the null state (0, 0, 0, 0, 0) does not change, and all other states lead to the state (1, 1, 1, 1, 1).

## Generation of Specified Cycle Lengths

In the case of a linear f.s.r. the null state (0, 0, 0, ....) can never change as no modulo-2 function of this state can produce a '1'. The maximum length cycle for a linear f.s.r. is thus $(2^n - 1)$, since there are $2^n$ possible states and one has been excluded. Apart from the null state, there can be no minor-cycles, since all other states lie on the major cycle.

This maximum length cycle can be generated with quite simple logic for reasonable values of $n$. Up to and including $n = 11$ (i.e. cycle length of 2047) only two feedback terms ($C_i$) are required in each case (with the exception of $n = 8$, cycle length 255 which requires four). The required feedback terms are given in Table 1.

**TABLE 1**

**Feedback for maximum length cycle linear f.s.r.**

*The table gives only those cases which have the smallest number of feedback connexions. Many other feedback connexions give maximum length (e.g. for n = 5, and feedback from 1st, 2nd, 4th and 5th stages).*

| $n$ | STAGES FROM WHICH FEEDBACK MUST BE TAKEN* | CYCLE LENGTH |
|---|---|---|
| 2 | 1st, 2nd | 3 |
| 3 | 1st, 3rd, or 2nd, 3rd. | 7 |
| 4 | 3rd, 4th, or 1st, 4th. | 15 |
| 5 | 3rd, 5th, or 2nd, 5th. | 31 |
| 6 | 5th, 6th. | 63 |
| 7 | 4th, 7th. | 127 |
| 8 | 4th, 5th, 6th, 8th. | 255 |
| 9 | 5th, 9th. | 511 |
| 10 | 7th, 10th. | 1023 |
| 11 | 9th, 11th. | 2047 |

** The '1's in the top row of the transition matrix.*

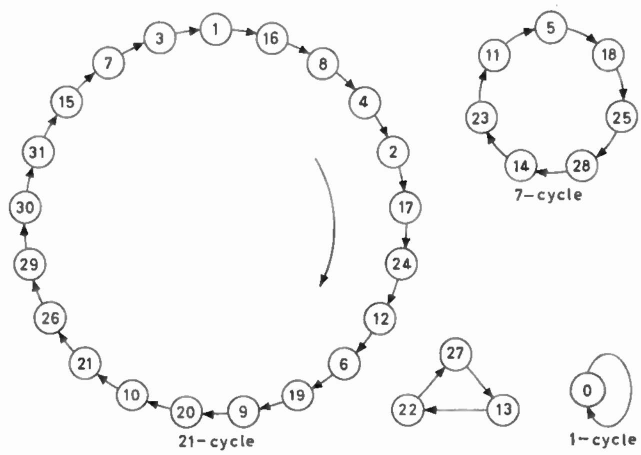


**Fig. 6. State diagrams for the circuit shown in Fig. 5**

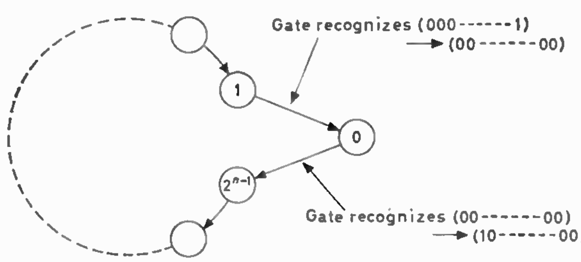


**Fig. 7. A state diagram including the 'null' state**

The unwanted null state can be avoided either by including an AND gate which recognizes the state and injects a '1' into the first stage, or by arranging that a switch-on transient in the equipment injects a '1' into one of the stages. Thus, cycles of 7, 15, 31, 65, 127 .... etc. are obtainable by this means.

It will be observed that the state (0000 .... 01) is always followed by (100 .... 00) with the above method. If an AND gate is included to recognize the state (000 .... 01) and convert it to the null state (0000 .... 0), and provided that a further AND gate has been added to move out of the null state, then operation on a cycle of length $2^n$ will occur. (Fig. 7).

A cycle of length $K$ where $k \leqslant 2^n - 1$ could be generated by a linear f.s.r. with different logic, but as explained above, it is only in the case of the maximum length cycle that unwanted minor cycles of various lengths can be avoided. A solution is described by Heath and Gribble[2] which involves designing a linear f.s.r. with cycle length $2^n - 1$, and introducing a 'jump' over certain states, to give a cycle of the required length $K$. (Fig. 8). It will be seen that as the original cycle was self starting, the new cycle must also be self starting. Bryant, Heath, and Killick[4] give a proof that for any $n$ and any $k =$

$2^n - 1$, there exists a unique state from which a jump can be made to obtain the desired cycle length.

The introduction of the 'jump' involves the addition of an extra $n$-input AND gate, to recognize the state from which the 'jump' is to be made and to change the state of the first stage. The determination of the 'jump' state for a given $k$ is relatively easy for small $n$, and Heath and Gribble[2] tabulate data for designing such f.s.r.'s for $k$ up to 127 (i.e. $n$ up to 7).

The feedback is no longer exclusively a modulo-2 function, so that the f.s.r. may now be reasonably termed 'non-linear'.

The foregoing methods enable an f.s.r. having any cycle length up to and including $2^n$ to be designed, and self-starting in the required cycle is guaranteed. However, the actual code generated is not open to choice. Thus, for $n = 3$, there are three different chain codes of length 6:

Only one of these would be produced by adding a 'jump' to a linear f.s.r. as described above.

In order to obtain a given code sequence which is

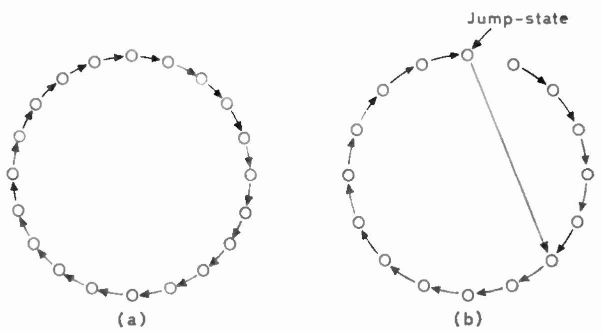


**Fig. 8. State diagrams with and without 'jump' state**
*(a) Original maximal cycle; (b) modified cycle with 'jump'*

different from that obtainable from a linear f.s.r. with a 'jump', it would be possible to add suitable extra logic (AND gate) to the outputs of the f.s.r. This is always possible, since any unique* $n$-digit code of cycle length $k$ can be produced from any other unique $n$-digit of the same length. For example, given any one of the codes given in Table 2, any of the others could be produced by appropriate logic. For example, given the sequence $(ABC) =$ 000, 100, 110, 111, 011, 001, this can be converted to $(DEF) =$ 100, 010, 101, 110, 011, 001, by the following logic:

**TABLE 2**

| (1) | (2) | (3) |
|---|---|---|
| 000 | 100 | 100 |
| 100 | 110 | 010 |
| 110 | 011 | 101 |
| 111 | 101 | 110 |
| 011 | 010 | 011 |
| 001 | 001 | 001 |

$$D = AB + A'B'C'$$
$$E = AB + AB'C'$$
$$F = A'C + ABC'$$

Thus, the possibility of generating any code of length $k$ ($k \leqslant 2^n$) has been shown. However, in many cases the procedure would not yield the simplest circuit.

An alternative is to abandon the linear f.s.r. altogether and to work entirely in terms of non-linear f.s.r.'s. The linear f.s.r. has the attraction of an elegant mathematical theory, which is at present entirely lacking in the case of the non-linear f.s.r. But the lack of a convenient mathematical theory does not mean that the non-linear f.s.r. also lacks practical advantages.

* *i.e., having no states repeated during $k$ successive states.*

### Practical Realization of Feedback Shift Register

Before going on to consider ways of designing non-linear f.s.r.'s it will be preferable to discuss practical realization in order to show the types of logic which lead to simple circuits. Referring to Fig. 1, the '0' state will be defined as a transistor switched 'off' and the '1' state a transistor switched 'on' thus, $A = 1$, $A' = 0$ denotes $VT_1$ switched 'on'. A potential of $V_{cc}$ volts is '0' and zero volts is '1'. The components $R_{A1}$, $R_{A2}$, $C_1$, $C_2$, $MR_1$, $MR_2$, ensure that each negative going trigger pulse is steered to the transistor which is 'on', so switching it 'off'. If $RB_1$ and $RB_2$ are included, and taken to $V_{cc}$ or 0 volts, then the circuit will change state whenever $(AP + A'Q) = 1$. If $P$ and $Q$ are not connected, then the

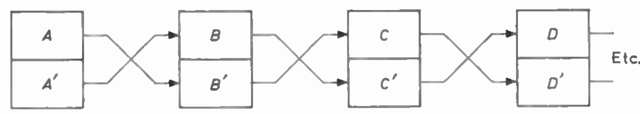


**Fig. 9. The ordinary shift register**

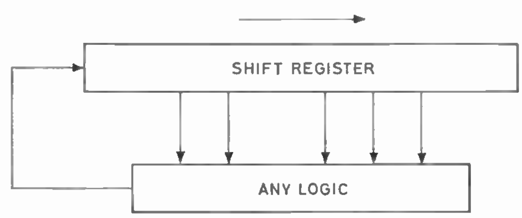


**Fig. 10. Arrangement of 'non-linear' feedback shift-register**

circuit changes state when $(A + A') = 1$. In other words, it *always* changes state, for every negative going trigger pulse.

Where $RB_1$ and $RB_2$ are included then it may not be obvious why $RA_1$ and $RA_2$ are also needed. However, if there exists any condition in the logic such that $PQ = 1$, then, without $RA_1$, $RA_2$, neither diodes would be reverse-biased, and a negative-going pulse would reach $VT_1$ and $VT_2$ simultaneously, resulting in a considerable increase in switching time, likely to cause errors in the switching sequence.

The presence of $RA_1$, $RA_2$ ensures that at least one diode is reverse-biased for any values of $P$ and $Q$, and thus ensures correct operation. However, it does mean that the amplitude of the trigger pulse must not be greater than $(V_{cc}/2)$.

Of course, if in a particular design, $P.Q$ is never '1', then these resistors may be left out. A common example where they are not necessary is the ordinary shift register (Fig. 9).

The circuit of Fig. 1 changes state whenever $AP + A'Q = 1$. This function will be referred to as $X$. The state of the stage after the change is then* $(AP' + A'Q)$, which is the function $S$ previously referred to. Thus, conversion from $X$ to $S$ is easily done, and it is often easier to work in terms of the function $X$. (Note that $A\,(+)\,X = S$, so $A\,(+)\,S = X$ and $S\,(+)\,X = A$).

* *Since* $S = A\ (+)\ X = AX' + A'X = A\ (AP + A'Q)' + A'\ (AP + A'Q)$
$= A\ (A' + P')\ (A + Q') + A'Q$
$= A\ (A'Q' + AP' + P'Q') + A'Q = AP'\ (1 + Q') + A'Q = (AP' + A'Q)$.

## Design of 'Non-Linear' f.s.r.

The case of Fig. 10 will be considered, in which no initial restriction is placed on the logic. From the 'change' signal $X = A'Q + AP$), it can be seen that the function $P$ and $Q$ should be as simple as possible, consisting ideally of a single variable.

Suppose one considers again the code:

| *ABC* | |
|---|---|
| 111 | 7 |
| 011 | 3 |
| 001 | 1 |
| 100 | 4 |
| 110 | 6 missing states (0, 2, 5) |

It can be seen that a possible function for $X$ would

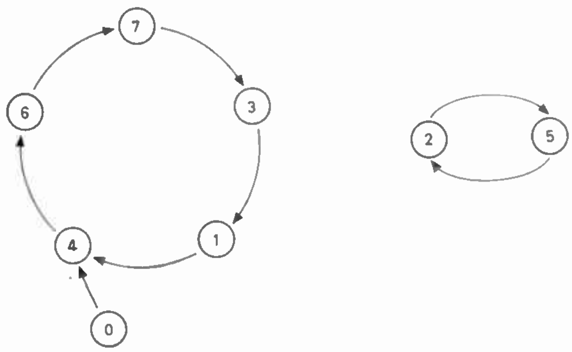

Fig. 11. State diagrams for a 'non-linear' feedback shift-register

(a)

| | *A* | *A'* |
|---|---|---|
| *B'C'* | 0 | |
| *B'C* | | 1 |
| *BC* | 1 | 0 |
| *BC'* | 0 | |

(b)

| | *A* | *A'* |
|---|---|---|
| *B'C'* | 0 | 1 |
| *B'C* | 1 | 1 |
| *BC* | 1 | 0 |
| *BC'* | 0 | 0 |

Fig. 12. A Karnaugh map
*(a) Original map; (b) map satisfying a, b and c*

be $X = (ABC + A'B'C)$. However, the arrangement would not be self-starting as the null state (000) cannot change.

An alternative function for $X$ would be $X = AC + A'(B' + BC')$ but, while ensuring that the (000) state leads into the major cycle, a new minor cycle (010, 101, 010, 101 ....) is possible (Fig. 11).

It is apparent that this procedure is not satisfactory as it stands, for a correct result cannot be guaranteed. In general, the states which cause difficulties are:

$\begin{Bmatrix} 1, 1, 1, 1, 1, \ldots \\ 0, 0, 0, \ldots\ldots \end{Bmatrix}$ these are likely to form cycles of unit length.

$\begin{Bmatrix} 1, 0, 1, 0, 1, \ldots \\ 0, 1, 0, 1, 0, \ldots \end{Bmatrix}$ these are liable to form a minor cycle of length 2.

To avoid such difficulties, it is necessary to ensure that, in addition to generating the required cycle, the function $X$ also satisfies the following conditions:

(a) $(A.B.C.D.E. \ldots.) = 1$

(b) $(A'B'C'D'E' \ldots.) = 1$

(c) $(AB\,CD\,E \ldots..)\ (A'B.C'DE' \ldots.) = 0$

For $n \leqslant 4$ an easy design procedure is to use the Karnaugh map for obtaining the simplest form for the function $X$. The map consists of a set of squares, one for each of the possible 'minterms'. (e.g. for $n = 3$, these are ($A'B'C'$, $A'B'C$, $A'BC$, $AB'C'$, $AB'C$, $ABC'$, $ABC$) and the Karnaugh map is shown in Fig. 12. From the code to be generated, the essential conditions for $X = 1$ and for $X = 0$ are inserted. (i.e. for the example, $ABC$ and $A'B'C = 1$, and $A'BC$, $AB'C'$ and $ABC' = 0$). Were it not for the likelihood of unwanted minor cycles, the remaining squares could be regarded as 'don't care' conditions, and filled with '1' or '0' in such a manner as to give the simplest function for $X$. To avoid the minor cycles, extra '1's must be added to satisfy conditions (a) and (b) (if not already satisfied). Some choice exists in satisfying condition (c), since $0{:}1 = 0$; $0{:}0 = 0$; and $1{:}0 = 0$. Finally, the remaining unfilled squares have '1's and '0's added in such a manner as to give the simplest $X$. In doing this, the aim is to obtain blocks of '1's together. In the example two blocks of two are obtained, giving the function $X = A'B' + AC$, (hence. $S = AC' + A'B'$). The circuit realization is shown symbolically in Fig. 13.

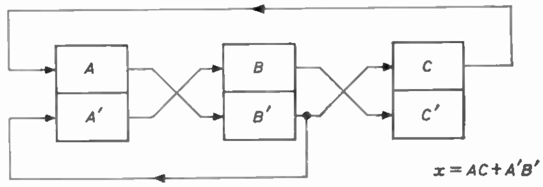

Fig. 13. The circuit realization for the state diagram shown in Fig. 11

As a further example, consider the generation of the following code:

| *ABCD* |
|---|
| 0000 |
| 1000 |
| 1100 |
| 1110 |
| 1111 |
| 0111 |
| 0011 |
| 0001 |

Fig. 14 shows the relevant Karnaugh map, first with the essential conditions to generate the cycle, and then with additional '1's added to give the simplest logic. To satisfy condition (c), both $A'BC'D$ and $AB'CD'$ are chosen to be 0. Thus, $X = AD + A'C'D'$, requiring an additional AND gate to form $C'D'$ (Fig. 15). The Karnaugh map is not easy to use for more than four variables, but it can be used for up to six.

## A More General Case

By taking feedback to other stages as well as the first, it is possible to generate code sequences which are not

Fig. 14. Karnaugh maps

| | *A'B'* | *A'B* | *AB* | *AB'* |
|---|---|---|---|---|
| *C'D'* | 1 | | 0 | 0 |
| *C'D* | 0 | | | |
| *CD* | 0 | 0 | 1 | |
| *CD'* | | | 0 | |

→

| | *A'B'* | *A'B* | *AB* | *AB'* |
|---|---|---|---|---|
| *C'D'* | 1 | 1 | 0 | 0 |
| *C'D* | 0 | 0 | 1 | 1 |
| *CD* | 0 | 0 | 1 | 1 |
| *CD'* | 0 | 0 | 0 | 0 |

$x = AD + A'C'D'$

chain codes. For example, the code:

$ABC$
1 1 1
0 1 0
0 0 1
1 0 1
0 0 0
1 1 0

could not be generated by any of the circuits discussed so far. But, by using the techniques of the preceding section, change signals $X_A$, $X_B$, $X_C$, can be found for the feedback to each stage.

$$X_A = AC + A'B'$$
$$X_B = BA' + B'C'$$
$$X_C = CA + C'B$$

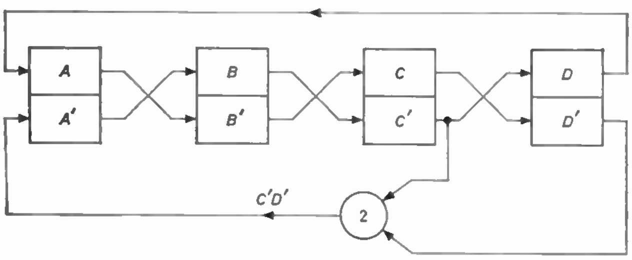


**Fig. 15. The circuit realization for the Karnaugh map shown in Fig. 14**

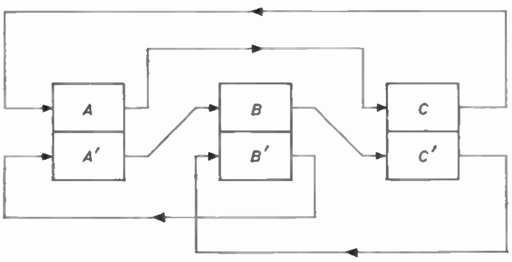


**Fig. 16. The circuit with feedback function to each stage**

A symbolic circuit realization is shown in Fig. 16. Such an arrangement can no longer strictly be called a shift register.

In general, the $X$ signal for the $R^{th}$ stage must be the form $X_R = R.P + R'Q$, (where $P$ and $Q$ are single variables if possible).

If the original code was a chain code, then the $X$ signals for all stages except the first will take the form of shift-register functions. (i.e. $X_B = B'A + BA'$, $X_C = CB' + C'B$, $X_D = DC' + CD'$, and so on). Thus, the f.s.r. with feedback only to the first stage, so generating chain codes, may be regarded as a special case of a more general arrangement with feedback to any stages, capable of generating any code.

Two more examples will be given for the generation of a code which is not a chain code.

(1) Generation of the code:

$ABC$
1 1 1
0 0 0
1 0 0
0 1 0
1 1 0
0 0 1
0 0 1
1 0 1
0 1 1

It is easy to see that $X_A = A + A'$ (i.e. a simple 'divide by two' circuit).

$$X_B = A = B'A + BA$$
$$X_C = AB = C(AB) + C'(AB)$$

The circuit realization is shown in Fig. 17.

(2) Generation of the code:

$ABCD$
0 0 0 0
0 0 1 0
0 1 0 0
0 1 1 0
1 0 0 0
1 0 1 0
1 1 0 0
1 1 1 0
1 1 0 1
1 0 1 1
1 0 0 1
0 1 1 1
0 1 0 1
0 0 1 1
0 0 0 1

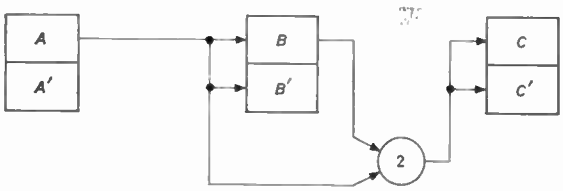


**Fig. 17. The circuit for a binary counter**

This code is of interest since digits $A$, $B$, $C$, count through each binary number from (000) to (111) and then back to (000). By means of the Karnaugh map method, the 'change' signals are found to be:

$$X_A = A(B'C'D) + A'(BCD')$$
$$X_B = B(A'D + AD') + B'(A'D + AD)$$
$$X_C = C + C'$$
$$X_D = D(A'B') + D'(AB)$$

## Conclusions

This article has described methods of designing cyclic-code generators, so that given any desired cycle-length or code-sequence, a circuit which produces it may be realized. An indication of the way to achieve the simplest circuit arrangement has been given.

In the case of the feedback shift register the output digit-sequence of the $R^{th}$ stage is identical to the output sequence of the $(R-1)^{th}$ stage apart from a one-digit delay. In the case of the general arrangement having feedback to each stage, the output sequence of each stage can be completely independent.

There are many possible applications for these circuits apart from frequency division and counting. They have been used by the authors for the direct generation of pulse code modulation signals for the testing of pulse code modulation receivers. Another application arises in frequency synthesizing equipment, for the generation of waveforms having a particular spectral distribution.


**Acknowledgment**

The authors thank the General Electric Co. Ltd for the permission to publish this article.


## APPENDIX

THE THEORY OF THE LINEAR F.S.R.

Suppose that the initial state of an $n$-stage f.s.r. is:

$$(X_1) = (A, B, C, D, E, \ldots) \quad \ldots\ldots\ldots \quad (1)$$

then the next stage will be:

$$(X_2) = (S, A, B, C, D, \ldots) \quad \ldots\ldots\ldots\ldots \quad (2)$$

where $S = C_1A\,(+)\,C_3C\,(+)\,C_4D\,(+)\,C_5E$ . . . .

and $C_1$, $C_2$, $C_3$ . . . . are all '1' or '0'.

Equations (1) and (2) may be conveniently written in matrix form, by regarding the states $(X_1)$ and $(X_2)$ as column vectors:

$$\begin{bmatrix} S \\ A \\ B \\ C \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} = \begin{bmatrix} C_1 & C_2 & C_3 & \cdots & C_{n-1} & C_n \\ 1 & 0 & 0 & \cdots & 0 & 0 \\ 0 & 1 & 0 & \cdots & 0 & 0 \\ 0 & 0 & 1 & \cdots & 0 & 0 \\ & \cdot & & & \cdot & \\ & \cdot & & & \cdot & \\ 0 & \cdots & & & \cdots & 0 \end{bmatrix} \begin{bmatrix} A \\ B \\ C \\ D \\ \cdot \\ \cdot \\ \cdot \end{bmatrix}$$

or more concisely:

$$(X_2) = (T)\ (X_1) \qquad (3)$$

The matrix $(T)$ is termed the 'transition matrix' of the f.s.r. Clearly, equation (3) can be extended to any state, so that:

$$(X_j) = (T)\ (X_{j-n}) = (T)^{j-1}(X_1)$$

The properties of the transition matrix determine completely the behaviour of the f.s.r.

For the case being considered, the matrix can always be partitional as follows:

$$(T) = \left[\begin{array}{c:c} C_1\ C_2\ \cdots\ C_{n-1} & C_n \\ \hdashline I_{n-1} & 0 \end{array}\right]$$

where $I_{n-1}$ is a unit sub-matrix of order $(n-1)$ by $(n-1)$. (In a more general case, with feedback to stages other than the first, the form of the matrix would not be so simple: however, the general case can always be decomposed into a direct sum of cases as above).

If the inverse matrix $(T)^{-1}$ exists, then each state has a unique predecessor, as can be seen by pre-multiplying both sides of equation (3) by $(T)^{-1}$.

$$(T)^{-1}\ (X_2) = (T)^{-1}\ (T)\ (X_1) = (I_n)\ (X_1)$$

$$\therefore\ (X_1) = (T)^{-1}\ (X_2) \qquad (4)$$

This applies to any two adjacent states, corresponding to reverse operation on the state diagram. If $(T)^{-1}$ exists, then equation (4) gives a unique result. If $(T)^{-1}$ does not exist, then the state diagram can have branches.

The condition for $(T)^{-1}$ to exist is that the determinant $[T]$ shall not be zero.

Evaluating $[T]$ by the rules of modulo-2 algebra gives $[T] = C_n$.

(Modulo-2 algebra is identical to conventional algebra with the exception that $1 + 1 = 0$ and $-1 = +1$. These exceptions make the algebra very simple).

Thus, the condition for $(T)^{-1}$ to exist is $C_n \neq 0$.

## Cyclic Behaviour

If $(T)^y = (I)$, then a cycle of length $y$ must exist, since $(T)^y\ (X) = (X)$ for any $(X)$. The major cycle has a length equal to the smallest integer $y$ satisfying the equation $(T)^y = (I)$.

This does not include the case of the minor cycles. The equation $(T)^x\ (X) = (X')$ implies $(T^x)\,(+)\,I)\ (X') = 0$, then there exist particular states $(X')$ for which $(T)^x\ (X') = (X')$. The length of minor cycles if any, are those values of $x$ for which the particular states $(X')$ exist. Thus, the condition for existence of a minor cycle of length $x$ is determinant $(T^x\,(+)\,I) = 0$.

The particular states $(X')$ can then be found by solving the equation $(T^x\,(+)\,I)\ (X') = 0$.

$(T)^y = (I)$ implies that $(T)^y\ (X) = (X)$ for any $(X)$ including those particular states $(X')$ forming minor cycles. It follows that the number of states $x$ in a minor cycle must be a factor of the number of states $y$ in the major cycle.

Major cycles having lengths equal to a prime number $Z$ in the range $2^n - 1 > Z \geqslant (2^{n-1})/3$ cannot be generated by a linear $n$-stage f.s.r.

For, suppose such a cycle could be generated. Then, the number of states not included in the cycle is $(2^n - Z)$. One of these is the null state, leaving $(2^n - 1 - Z)$ to form minor cycles. But, $Z$ is prime, and so has no factors, and such minor cycles cannot exist, and there can be no such states. Either $(2^n - 1 - Z) = 0$. $\therefore\ 2^n - 1 = Z$ which is contrary to the original supposition, or $Z$ is itself a factor of $(2^n - 1 - Z)$ i.e. $(2^n - 1 - Z) = 2Z$. $\therefore\ 2^{n-1}/3$. Thus, the original statement is verified.

## Characteristic Polynomial

The cyclic behaviour of the f.s.r. may be determined from the 'characteristic polynomial' of $(T)$ which is determinant $[T\,(+)\,I] = \lambda^n + C_1\lambda^{n-1} + C_2\lambda^{n-2}$ . . . . . $C_n$.

For the f.s.r. to yield a 'maximal length' cycle of $(2^n - 1)$ states, these polynomials are tested by Elspas[1] and Petersen[2] for many values of $n$.

From any matrix yielding a maximal-length cycle, it is possible to derive a second matrix which also yields a maximal-length cycle but with a different ordering of the states forming the cycle. For, suppose $[T_A]$ is one such matrix. Then $[T_A]^{-1}$ must exist (as there can be no branches) and this matrix, regarded as a new transition matrix, must also give a maximal-length cycle. This inverse matrix has the form:

$$[T_A]^{-1} = \begin{bmatrix} 010000 \cdots\cdots 0 \\ 0010000 \cdots\cdots 0 \\ 00010 \cdots\cdots\cdots 0 \\ \cdots\cdots\cdots\cdots \\ \cdots\cdots\cdots\cdots \\ 1\ C_1C_2C_3\ \cdots\ C_{n-1} \end{bmatrix} = \left[\begin{array}{c:c} 0 & I_{n-1} \\ \hdashline 1 & C_1C_2C_3 \cdots C_{n-1} \end{array}\right]$$

The feedback is now to the last stage, with shifting from right to lift instead of left to right, but this is merely equivalent to a re-labelling of the stages in the reverse order. For the normal direction of shifting, the matrix becomes:

$$[T_B] = \left[\begin{array}{c:c} C_{n-1}\,C_{n-2}\,C_{n-3} \cdots C_2C_1 & 1 \\ \hdashline I_{n-1} & 0 \end{array}\right]$$

The original matrix $[T_A]$ was:

$$[T_A] = \left[\begin{array}{c:c} C_1\,C_2\,C_3 \cdots C_{n-1} & 1 \\ \hdashline I_{n-1} & 0 \end{array}\right]$$

Thus, $[T_A]$ and $[T_B]$ differ only in that the order of the first $(n-1)$ terms in the top row is reversed. ($C_1C_2C_3$ . . . . . . $C_{n-1}C_n$) becomes ($C_{n-1}C_{n-2}$ . . . . . $C_3C_2C_1C_n$), and $C_n = 1$ in all cases because $[T_A]^{-1}$ must exist. For the maximal-length cycle cases $[T_A]$ and $[T_B]$ are never identical* (see Reference 3, p. 106).

For cases which are not maximal-length, but for which $c_n = 1$, the above reasoning still applies (though occasionally $[T_A]$ and $[T_B]$ may be identical) so at least two realizations of any particular cycle-pattern are possible.

The characteristic polynomials of $[T_A]$ and $[T_B]$ are $\lambda^n\,(+)\,C_1\lambda^{n-1}\,(+)\,C_2\lambda^{n-2}\,(+)$ . . . . . $(+)\,C_{n-1}\lambda\,(+)\,C_n$ and $C_n\lambda^n\,(+)\,C_{n-1}\lambda^{n-1}\,(+)\,C_{n-2}\lambda^{n-2}\,(+)$ . . . . . $(+)\,C_1\,(+)\,1$ respectively, these being reciprocals of one another†. The significance of the case where $C_n = 0$ can now be seen;

* *Except for the trivial cases with $n = 1$ and $n = 2$, for which $[T_A]$ has for its top row [1] and [1,1] respectively.*

† *The reciprocal polynomial $F^*(\lambda)$ of a given polynomial $F(\lambda)$ is defined by $F^*(\lambda) = \lambda^n F(1/\lambda)$, where $n$ is the degree of $F(\lambda)$.*

if $[T_A]$ was the transition matrix of $n$-stage f.s.r. (having an $n^{th}$ degree characteristic polynomial) then the reciprocal polynomial would have degree $(n-1)$ if $C_n = 0$, and so could not relate to an $n$-stage f.s.r. and no corresponding $[T_B]$ could exist.

*Example*

Consider the five-stage linear f.s.r. with $S = D\,(+)\,E$ referred to previously.

The transition matrix is thus:

$$[T] = \begin{bmatrix} 00011 \\ 10000 \\ 01000 \\ 00100 \\ 00010 \end{bmatrix}$$

$C_5 = 1$ so $[T] = 1$, and therefore $[T]^{-1}$ exists and there are no branches on the state diagram.

The characteristic polynomial is $\lambda^5 (+) \lambda (+) 1$ which is not one of the irreducible polynomials, and so a maximal-length cycle is not produced.

If $[X_1]$ is $(0, 0, 0, 0, 1)$ then $[X_2] = [T]\ [X_1]$

$$\therefore\ [X_2] = \begin{bmatrix} 00011 \\ 10000 \\ 01000 \\ 00100 \\ 00010 \end{bmatrix}\begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 1 \end{bmatrix} = \begin{bmatrix} 1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$

and so on.

Since the characteristic polynomial is $\lambda^5 (+) \lambda (+) 1 = 0$, and since every matrix satisfies its own characteristic polynomial (Cayley's theorem), then $[T]^5\ (+)\ [T]\ (+)\ [I] = 0$

$$\therefore\ [T]\ [T^4 (+) I] = [I]$$

Taking the fourth power of the original equation:

$$[T]^{20} (+) [T]^4 = [I] \qquad \therefore\ [T]^4 (+) [I] = [T]^{20}$$

$$\therefore\ [T]\ [T]^{20} = [I]$$

$$[T]^{21} = [I]$$

Thus $[T]^{21} [X] = [X]$ for any $[X]$ and the length of the major cycle is 21.

Determinant $[T^3 (+) I] = 0$ and determinant $[T^7 + I] = 0$, so that minor cycles of length 3 and 7 exist. (Note that 3 and 7 are factors of 21).

For the minor cycle of length 3.

$$[T]^3 = \begin{bmatrix} 01100 \\ 00110 \\ 00011 \\ 10000 \\ 01000 \end{bmatrix} \qquad \therefore\ [T^3 + I] = \begin{bmatrix} 11100 \\ 01110 \\ 00111 \\ 10010 \\ 01001 \end{bmatrix}$$

Then, the particular states forming this minor cycle are given by:

$$[T^3 (+) I] \begin{bmatrix} A \\ B \\ C \\ D \\ E \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$

$$\therefore\ A = D \text{ and } B = E \text{ and } A\,(+)\,B\,(+)\,C = 0$$

so either $A = D = B = E = 1$, and $C = 0$
or $A = D = B = E = 0$, and $C = 0$
or $A = D = 1$, $B = E = 0$, and $C = 1$
or $A = D = 0$, $B = E = 1$, and $C = 1$

then excluding the null state, these states are:

$(ABCDE) = (11011), (10101), (01101)$, which are the three states of the minor cycle.

As $[T]^{-1}$ exists, then another matrix giving the same cyclic pattern can be obtained by interchanging the first four terms of the top row:

$$[T]' = \begin{bmatrix} 10001 \\ 10000 \\ 01000 \\ 00100 \\ 00010 \end{bmatrix}$$

This matrix will be found to give a major cycle of length 21, and minor cycle of lengths 3 and 7, but the states forming the cycles differ from those of Fig. 5. Thus, for the original matrix the cycle of length 7 was formed from states 14, 23, 11, 5, 18, 25, 28, while for the new matrix the states are 26, 29, 14, 7, 19, 9, 20.

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## Remote Control by Telephone

A new development of G.E.C. (Electronics) Ltd is the introduction of an automatic answering unit for use in conjunction with the Company's range of Teledata frequency multiplex equipment.

With this new equipment, a wide range of industrial plant can be controlled by telephone over normal G.P.O. exchange lines. From control centres anywhere in the country, engineers can rapidly control and supervise remote plant such as isolated and unmanned oil, gas, water or sewage pipeline pumping stations.

The same equipment can be used to obtain information on the operating state of the remote plant.

The first of a number of such systems has now been installed in England on behalf of Shellmex and B.P. Pipelines Department to operate motorized valves at isolated points along an oil pipeline. Hitherto, staff have had to travel long distances daily to adjust the valves by hand; adjustments can now be made for the cost of a telephone call. Significantly, also, emergency action can be taken within seconds using the new system.

Normally up to ten supervisory channels are offered, but, by time-sharing, upwards of 50 channels can be provided. Installation of the equipment does not interfere with the normal use of the telephone at the distant station. To control or interrogate a remote station, the engineer dials the appropriate telephone number. After a ringing period of 10sec , the remote station announces its identity by means of a message recorded on a magnetic drum. The engineer then transmits a 'hold' signal to maintain the link, and the condition (on/off; open/shut) of the valves at the station is automatically indicated to the control centre. Changes in the station's condition are brought about by operating a control key; automatic confirmation is given when the required changes have been completed. The circuit is then cleared by releasing the 'hold' key.

The system is safeguarded against unauthorized operation by spurious signals. If the station is interrogated in error, it will announce its identity, but will clear the line if the 'hold' signal is not transmitted.

A built-in sealed battery ensures 'fail-safe' operation. In the event of failure of mains supply to the station all functional relays are held at their latest settings until power is restored.

The equipment operates in ambient temperatures of from −20°C to 70°C. A unit form of construction is used so that a wide range of industrial applications can be met by the basic equipment.

Other obvious applications for the equipment are the remote control of electricity sub-station circuit-breakers and transformer tappings and the start-up of remote emergency stand-by plant. Remote control of radio transmitters is also possible, including frequency changing and aerial switching.

# Stabilizer Tube Test Set

By D. T. Smith*, B.Sc., A.M.Brit.I.R.E.

*This article describes a test set for voltage stabilizer and reference tubes, and Zener diodes. Currents in the range 0·5 to 20mA are passed through the device under test, and the voltage developed measured on a voltmeter.*

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

THE test set described in this article was built for use in a laboratory where a wide range of electronic equipment is used. During the maintenance of this equipment it is often necessary to check valves, and this test set will check stabilizer and reference tubes that are not covered by normal valve testers. It has been made versatile to cover a wide range of tubes and also Zener diodes.

To test a stabilizer tube it is necessary to pass a known current through it, and monitor the voltage developed. As this voltage should vary only slightly with changes in current, it is not necessary to have a precise control of current. However, it is necessary to be able to control the current to the correct working region for the device. Measurement of the voltage, however, should be reasonably accurate.

* *Clarendon Laboratory, University of Oxford.*

## Circuit Description

The circuit consists basically of a constant current source feeding into the device under test, with a voltmeter measuring the voltage developed across the device. A pentode valve $V_3$ is used as the current source, with the device under test connected between anode and h.t.+; the grids being held at constant potentials. A stabilizer tube $V_2$ holds the screen grid at about 150V above h.t.−, and a potential divider holds the control grid at about 20V. The current through the valve is thus determined primarily by the value of cathode resistance, and this can be switched to give currents in the range 0·5 to 20mA in steps of ratio 2:1. There is in fact some variation in output current as the output voltage is varied over the working range, but it remains within $\pm 10$ per cent of nominal. The voltmeter has ranges from 5 to 200V f.s.d. and is fully protected against overloads by the Zener diode $MRZ_2$ which limits the current to less than twice the f.s.d. current.

Connexions to the device under test are made either via terminals or via pin selector switches to one of the selection of valve holders. A terminal and selector switch are included for priming or ignition electrodes, and when these are used a suitable resistor is connected between anode and priming terminals. If precision voltage measurements are required, an external precision voltmeter may be connected to the anode and cathode terminals.

To enable some dynamic measurements to be made when required, a small alternating current is also fed into the device. This is approximately a square wave at 50c/s of 100$\mu$A peak-to-peak, stabilized in amplitude by the Zener diode $MRZ_1$. The resulting alternating voltage may be monitored by connecting an oscilloscope between the anode and cathode terminals.

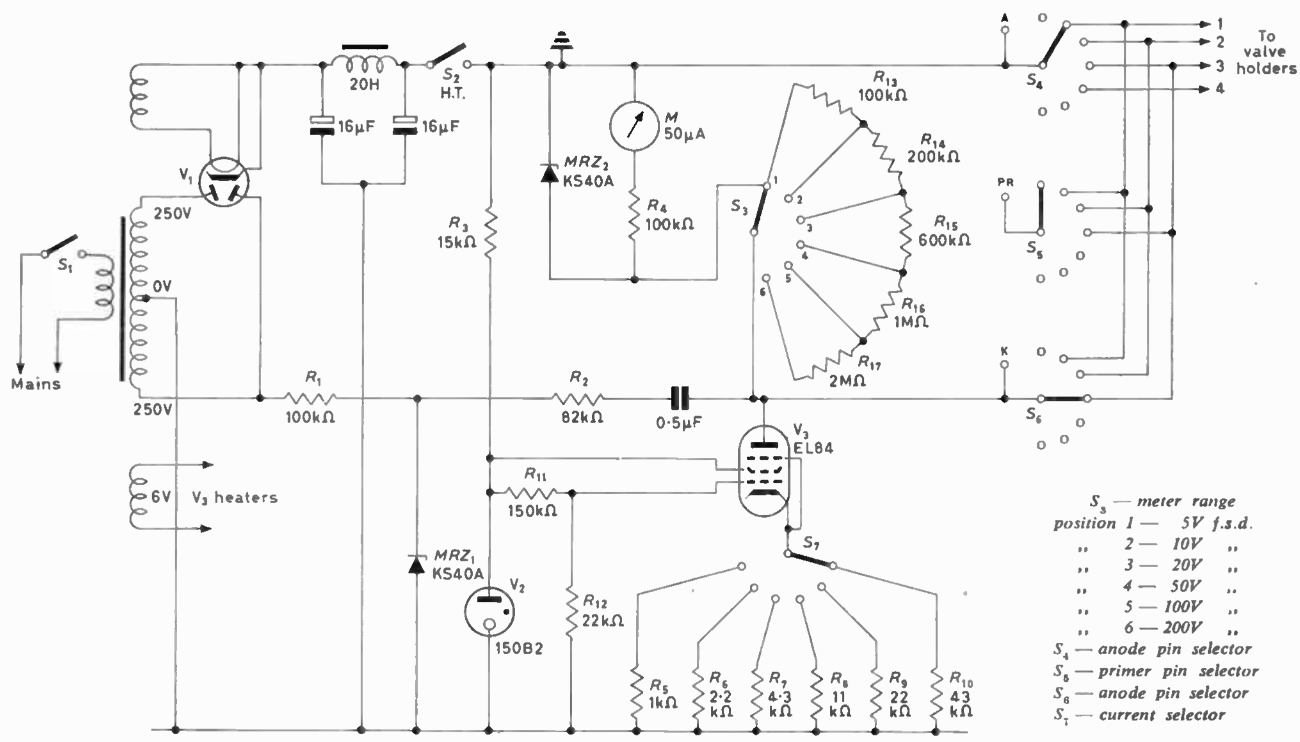



**Fig. 1. Stabilizer tube test set**

# BOOK REVIEWS

### Fundamentals of Microwave Electronics

**By V. N. Shevchik. 253 pp. Med. 8vo. Pergamon Press. 1963. Price 70s.**

THIS book is the sixteenth volume in the International Series of Monographs on Electronics and Instrumentation, a series which includes books such as that on probability and information theory as applied to radar by Woodward, on space charge waves and slow electromagnetic waves by Beck and on environmental testing by Dummer and Griffin. It must be said at once that, judged by these standards, or indeed by any reasonable standards, the book is not a success: one can only conclude that the publishers have been more concerned to preserve the international character of their series than its technical value.

The main defect of the book is that it is seriously out of date. It is based on a course of lectures given in the Radiophysical Department of Saratov State University (Saratov is a city on the Volga, almost due north of Stalingrad), and is said to give a detailed account of the subject up to 1957.

As is admitted in the foreword to the English edition, a number of important advances which have occurred since 1957 are not recorded. This deficiency is made up in part by the foreword, which covers masers, parametric amplifiers, very low noise travelling-wave tubes and tunnel diodes in five pages and various other devices in another five, with a bibliography of some 50 entries, almost all American, only one being dated as late as 1961. One must rate this foreword as a brave attempt at the impossible task of making the book only three years out of date.

The main body of the book opens with an author's foreword and an introduction: then come parts one (one chapter on oscillatory circuits, 20 pages), two (two chapters, totalling 90 pages, on fundamental electronic phenomenon and on energy exchange between electron beams and electric fields) and three (four chapters of 130 pages on various microwave electronic devices). There follows a bibliography and an index: it is interesting to note that the bibliography to chapter 1 has fifteen references, of which eleven are Russian (three of them translations from the English—surely references to the English editions would have been more useful?) whereas that to chapter 6 has three Russian out of 26, including 'Travelling Wave Tubes' translated from the English. This is presumably J. R. Pierce's famous book. Pierce, who might be called the father of post-war electron beam science, is mentioned only once in the bibliography and is wrongly initialled in the index.

It may be asked whether the book, despite its obsolescence and the grave handicap of an unfamiliar notation, has sufficient clarity or freshness of viewpoint to redeem these defects. Clarity is not helped, for instance, by the muddle between text and diagrams on pages 8 and 9: and the coupled mode theory of the interaction between beams and circuit voltages, which gives probably the best physical picture available of the action of all microwave tubes, has no place in the text. A. H. W. Beck's book on space charge waves and slow electromagnetic waves, first published in 1958, and number 8 in the same International Series, devotes considerable space to the coupled mode theory, and is surely a far better account of the field than the book now under review.

M. J. B. SCANLAN

### Theory of Automatic Control

**By M. A. Aizerman. 519 pp. Med. 8vo. Pergamon Press. 1963. Price £4**

READING translated Russian technical books sometimes provokes a state of mind analogous to Alice's experiences in 'Through the Looking Glass', and this book does just that. The situation is familiar enough, but all the names of the heroes are different and the story has unexpected twists. Only the beginning and the end are as we know things.

The text commences with a functional description of industrial controllers of various kinds including continuous, on-off, and extremal controllers. Chapters 2, 3, and 4 deal at length with the analysis of linear systems, including the technique of developing linear mathematical models, the stability of these models, and their response to random inputs. Lastly, chapter 5 has the title 'Auto and forced oscillations in non-linear systems', and provides a thorough introduction to non-linear control system theories and graphical methods. Two appendices summarize Laplace and Fourier Transform theory and gives the numerical values of some useful trigonometrical functions. Finally there is an apparently excellent list of references of Russian work, some, but not all, of which may be obtainable here.

All the proper ground is covered, but sometimes using unfamiliar concepts like the so-called D-partition method of determining stability boundaries. Since such concepts are just another approach, and we already have Nyquist and Dzung *et al*, some editorial notes would have been of value; for example, broadly to equate Mikhailov with Nyquist, perhaps. This would probably help the reader, as would a slightly less stilted translation here and there. Nevertheless, the translators have done an excellent job on the main text, despite an inexplicably uncorrelated index.

That a work of such high standard was worth translating is indisputable.

K. C. GARNER

### Electromagnetic Fields

**By S. A. Schelkunoff. 413 pp. Med. 8vo. Blaisdell Publishing Co. 1963.**

PROFESSOR SCHELKUNOFF'S book is written to cover two courses in electromagnetic field theory. Chapters 1 to 5 constitute the first course, which would be suitable for degree students. The final four chapters cover more advanced work.

The book is different, in many respects, from other similar texts and is clearly the work of an author experienced in the application of his subject. At the first reading, the reviewer found the order in which the subject matter was presented a little confusing, but further examination showed it to be quite logical.

Chapter 1, dealing with "Basic Concepts and Equations" covers a wide range of subjects from force, mass, work, etc., to Maxwell's equations and boundary conditions. Then Chapter 2, "Static and Almost Static Fields" starts with potential, charge distributions, etc., and progresses to transmission lines. Chapter 3 is concerned with energy, and starting with stroage leads to further study of transmission lines, including equivalent circuits. In Chapter 4, "Waves", the transmission line equations are derived from Maxwell's equations and then various important formulae are deduced, including a discussion of skin effect. An approximate solution is then obtained to the problem of wave propagation in hollow tubes of both rectangular and circular cross-sections. This approach is an example of the author's great stress on the power and application of approximate methods of solution, which are important in many cases for which an exact solution is not possible.

Chapter 5, "Spherical Waves", is concerned mainly with radiation fields; "Normal Modes", Chapter 6 are introduced by reference to d.c. examples, the current distribution in conducting plates of varying cross-section being studied. Chapter 7 concerns reflection and scattering of both plane and guided waves. The last chapter deals with "Coupled Oscillations" and coupled modes including transmission line directional couplers.

The book concludes with 90 pages of problems, with solutions. These expand many points touched upon in the main body of the book where the treatment is necessarily concise as so much ground is covered in a single volume of this size. It is felt that most engineers, interested in this subject, would gain new insights by a study of this work.

P. C. BUTSON

## Multilinear Analysis for Students in Engineering and Science

**By G. A. Hawkins. 220 pp. Demy 8vo. John Wiley & Sons Ltd. 1963. Price 49s.**

MULTILINEAR analysis is tensor analysis and the author was led to write this book to help students of the theory of continuous media, but the book should be valuable to many engineers. The publication of several books of this type suggests that there are not only many engineers wishing to understand tensor analysis, but many more who need to use it without fully understanding the subject.

This book is admirable for reference as it contains all the useful formulae in detail and several valuable tables. For those who find this subject very difficult to understand it has the merit of very detailed proofs step by step—this is the only book known where two matrices are multiplied in complete detail, a boon to the lecturer. The presentation and diagrams are clear, and the introduction to the notation and structure of tensors are excellent.

The subject is presented as it is used in practice, as a method of handling equations in such a form that they are valid in all systems of co-ordinates.

The book was, however, found a little dull. This may be because there is little sparkle in the proofs, but it is thought that it is due to the lack of illustrations of the use of tensor methods, to provide a systematic analysis of equations to be solved. The brief chapter on the inertia tensor is simply a proof that there is an inertia tensor: the idea of principal axes is not mentioned anywhere. Many books from the U.S.A. share this fault, however, and there may be many reasons for it.

It is certain that many will buy and use the book and the author will have many irate letters about the ridiculous diagram on page 79.

G. J. KYNCH

## The Physics of Magnetic Recording

**By D. C. Mee. 270 pp. Crown 8vo. North-Holland Publishing Co., Amsterdam. 1964. Price 60s.**

Volume II in the series "Selected Topics in Solid State Physics", edited by E. P. Wohlfarth.

THE aim of this book is to describe the techniques of magnetic recording from the point of view of the physicist. The subject is treated on an intermediate level and more use is made of simple physical models than of rigorous mathematical derivations.

Chapter I is introductory and is used to explain the scope of the text which covers not only moving media magnetic storage devices but static magnetic memories used for limited storage in computers, where the speed of information retrieval must be high.

Chapter II deals comprehensively with the magnetic recording process when a.c. bias is used. It details first a simple model to illustrate anhysteretic magnetization and then discusses the various ways in which a practical a.c. bias system must differ from the model.

Chapter III covers zero and unidirectional bias methods and pulse recording. It has special interest for both video-tape and computer engineers.

Chapter IV is a comprehensive treatment of various magnetic reproducing processes and it covers head design, background and modulation noises and print-through.

Chapters V and VI deal with tape design criteria and the properties and preparation of various types of coatings. An interesting section covers magnetic measurement techniques.

Chapter VII looks at various recording techniques and discusses the design considerations for a wide-band, high-resolution recording system. Finally, equalization problems are investigated.

This 270-page book is clearly printed and admirably illustrated and there is a comprehensive bibliography and a carefully compiled index.

Dr. D. C. Mee has adroitly marshalled his facts and his approach, though rather puzzling in places, is thought-provoking. As he admits, there is still a lot of important work to be done in this field but the machinery of anhysteretic magnetization is beginning to be better understood. On the trend of future developments Dr. Mee is content merely to indicate some of the ways in which development may be expected.

To sum up, this book is a must for any electronic engineer interested in the physics of magnetic recording. It will also provide interesting reading for physicists and chemists who, as the editor remarks in his preface, will find in this book that their basic researches need not remain in the realm of pure science forever, but will gain vitality by cross-fertilization with technological application. This book should therefore have a wide appeal.

P. J. GUY.

## The Analysis of Linear Systems

**By W. H. Chen. 577 pp. Med. 8vo. McGraw-Hill, New York. 1963. Price 124s.**

THE author is Professor of Electrical Engineering at the University of Florida, and his aim is to provide a systematic discussion of the important principles, concepts and techniques required, suitable for advanced undergraduates or graduate level. Only lumped-element systems are considered.

The coverage is comprehensive—Analysis by loops and nodes, by matrices, by means of topology, Fourier series, the principle of superposition, signal-flow diagrams and other topics are all fully discussed. Treatment is mainly based on the Laplace transform. The last part deals with stability, and includes both the Routh and the Nyquist criteria. The diagrams are very clear and there do not appear to be any serious misprints.

As a work of reference, therefore, the book is very useful. But the very comprehensiveness which is an advantage from the point of view of reference is a disadvantage from the point of view of a student, graduate or practising enginer who wants to start from scratch and grasp the essential principles and to be able to apply them to reasonably straightforward practical problems unaided. For such a person, the variety of possible techniques is confusing, and a way of understanding the essentials without being confused by detail is the prime requirement. Matrices (especially those associated with four-terminal networks) provide an outstanding means of determining the relations between a circuit and the external world without the necessity of bothering about unwanted internal currents and voltages. Such matrices are discussed, but this important advantage is inadequately stressed. Unfashionably, the reviewer regards operational calculus as simpler and more general than Laplace-transform calculus, and therefore regrets that operational calculus is barely mentioned. Again, in the matter of stability criteria, Routh's rule that for the characteristic equation

$$s(s^2+\alpha_1^2)(s^2+\alpha_2^2)\ldots(s^2+\alpha_r^2)+(s^2+\beta_1^2)(s^2+\beta_2^2)\ldots(s^2+\beta_r^2)=0$$

all the quantities $\alpha$ and $\beta$ must be real and must interlace is much easier to apply than is often realized, especially for characteristic equations of degree not exceeding 7.

Thus while the book can be commended as a work of reference to someone who is already knowledgeable and wishes to refresh his memory on specific topics, the book seems to lack the transversal view that can permit anyone starting *ab initio* to grasp first the basic, simple essentials, and then build up for himself a sound technique from a sure foundation.

J. W. HEAD

### How Television Works

**By W. A. Holm. 351 pp. Demy 8vo. 2nd Edition. Philips Technical Library. 1963. Price 35s.**

It is a completely revised edition of an earlier work and opportunity has been taken to enlarge and extend the scope of various sections and bring the whole of the text right up to date. In particular, the chapters on amplifiers, studio techniques and large screen projection have been considerably expanded.

### Colour Television Explained

**By W. A. Holm. 110 pp. Demy 8vo. Philips Technical Library. 1963. Price 20s.**

The aim of this book is to describe in detail the theoretical basis of colour television.

### Problems of the Design and Accuracy of Complex Continuous Action Devices and Computer Mechanisms

**Edited by N. G. Bruyevich. 264 pp. Med. 8vo. Pergamon Press. 1964. Price 70s.**

This book consists of papers dealing with the design of mechanical and electromechanical analogue computers and contributed by leading Russian authorities in this subject.

The book is divided into two parts. The first part contains papers on problems of the design and accuracy of continuous action complex devices. The term 'devices' includes mechanisms and electrical circuits of a machine instrument.

The second part of book is called 'Accuracy of some typical computing devices'. In this part there are seven articles, each devoted to the study of the accuracy of some actual simple mechanism.

### Transistor Amplifiers for Audio Frequencies

**By T. Roddam. 252 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 45s.**

This is a practical book on the design of audio-frequency transistor amplifiers.

Although this is an introductory work intended primarily for those new to the subject, experienced designers will find there is much that they can learn from it. The book is readable and the mathematics have been kept as simple as possible.

The author starts with the physics of transistor action. Then follow, chapters on small signal behaviour, representation of equivalent circuits and the large signal behaviour of transistors.

Next, the author deals with Class-A circuits; biasing; interstage and decoupling methods; and multi-stage Class-A amplifiers, followed by a chapter on Class-B operation. The final chapter covers frequency-selective amplifiers and negative impedance.

### Principles of Feedback Design

**By G. Edwin and T. Roddam. 238 pp. Demy 8vo. Iliffe Books Ltd. 1964. Price 45s.**

The book commences with an examination of the application of negative feedback to simple amplifiers, of both the valve and transistor types.

The second half of the book deals with more general problems including signal flow diagrams, a discussion of the analytical approach and the use of feedback amplifiers as filters. The final chapter discusses in broad outline a miscellany of other feedback problems with the object of showing how a knowledge of basic principles can assist in the understanding of the behaviour of a wide range of closed loop systems.

### Understanding Television

**By J. R. Davies. 504 pp. Demy 8vo. Data Publications Ltd. 1964. Price 37s. 6d.**

This book deals in full with 405 and 625 line systems as well as the basic aspects of modern colour reception and reproduction.

Each new subject is introduced in terms which have been made clear in earlier chapters, and the reader is required to have only a basic knowledge of elementary radio principles. The treatment is non-mathematical.

### Transistor Electronics in Instrument Technology

**By N. I. Chistyakov. 378 pp. Demy 8vo. 1964. Pergamon Press. 1964. Price 80s.**

The papers contained in this volume review many developments achieved by Soviet research institutes, and academic and industrial laboratories.

The papers were presented at the first conference organized by the Scientific Technical Society at Moscow, in December 1956. The meeting was devoted to the theory, calculation, physical principles of operation. construction and technical applications of the most important semiconductor photocells, galvanomagnetic sensing elements, thermocouples, cooling elements, diodes and transistors.

Among the subjects covered are the problems of determining the amplifying parameters of semiconductor triodes and the operation of transistors in amplifier circuits. the theory and applications of transistor oscillators for sinusoidal oscillations. the practical application of transistor amplifiers in computers. automation and remote control telemetering.

### Wireless and Electrical Trader Year Book 1964

**448 pp. Demy 8vo. 35th Edition. Iliffe Books Ltd. 1964. Price 27s. 6d.**

Wireless and Electrical Trader Year Book, which was first published in 1925, has become an important reference book to the radio and electrical industries. It is the standard guide for all connected with sales or services, and of assistance to overseas buyers wanting to contact British sources of supply.

### Modern Dictionary of Electronics

**Compiled by R. F. Graf. 370 pp. Med. 8vo. W. Foulsham & Co. Ltd. 1963. Price 45s.**

The book contains over 10 000 electronic terms and definitions together with a specially written chapter for the guidance of the English reader by W. Oliver and has a pronunciation guide.

# Short News Items

**The Electronics, Science and General Divisions** of the Institution of Electrical Engineers are arranging a conference on "Lasers and their applications" at the IEE Headquarters on 29 September to 1 October 1964.

The aim of the conference is to survey present knowledge of lasers, coherent-light and quasi-coherent light sources and the possible applications of lasers, as a guide to potential users and those concerned with device developments.

The conference will cover:

Gas, solid-state (injection) and liquid type lasers.

Ancillary devices, including optical superheterodyne receivers, detectors and modulators.

Propagation, in free space and optical pipes.

Techniques for power, noise, spectrum and coherence measurements.

Optical radiation hazards, including physiological effects and necessary precautions.

Applications, including communications, ranging and navigation, micromachining and microwelding, and uses in medicine.

The material for the conference will take two forms: main survey contributions of up to 3 000 words and supporting contributions relating to a specialized aspect or new development of up to 700 words.

Those wishing to submit material are invited to send the suggested title and a synopsis of about 200 words to the Secretary of the Institution as soon as possible.

Further details and registration forms may be obtained from the Secretary, the Institution of Electrical Engineers, Savoy Place, London, W.C.2.

**The Institute of Physics and The Physical Society** announces that its Electronics Group is arranging a Conference on Electron Emission to be held in the University of Keele on 1 to 2 October, 1964. The main emphasis of the meeting will be on secondary electron, photo, and field emission phenomena, and also on recent developments in other fields of electron emission; review papers covering these topics have already been invited. Abstracts of about 200 words (submitted in accordance with the guidance given in leaflet ACB78 obtainable from the Institution and Society), should be submitted to Dr. A. K. Jonscher, Chelsea College of Science and Technology, Manresa Road, Chelsea, London, S.W.3, as soon as possible but not later than 1 June, 1964.

During the Conference Professor Martin Ryle, F.R.S., Professor of Radio Astronomy in the University of Cambridge, will deliver the 1964 Guthrie lecture.

Residential accommodation will be available in University Halls of Residence. Advance registration for attendance at the Conference is necessary. Application forms and further details will be available later from the Administration Assistant, The Institute of Physics and The Physical Society, 47, Belgrave Square, London, S.W.1.

**The Sixth Annual Conference** of the ASLIB Electronics Group is to be held at Ashridge College Berkhamsted on 5 to 7 June, 1964.

There will be two main sessions, one on "Computers, Mechanization and Documentation" in which Dr. M. P. Barnett, Computer Laboratory, M.I.T. and Mr. H. F. Vessey, T.I.L., Ministry of Aviation will speak, and one on "Indexing and Information Retrieval" in which KWIC, U.D.C., Co-ordinate Indexing, Facet and Alphabetical Systems will be discussed in a series of five short papers by users of each system.

Further details are available from the Conference Secretary Mr. R. S. Lawrie, Chief Librarian, Sperry Gyroscope Company Limited. Downshire Way, Bracknell. Berks.

**The Sectional Meeting of the World Power Conference** is to be held in Switzerland at Lausanne on 13 to 17 September 1964 and the theme selected for the meeting will be "The Struggle against Losses in the Field of Energy Economics".

Further details are available from the Secretary, British National Committee, World Power Conference, 201-2, Grand Buildings, Trafalgar Square, London, W.C.2.

**The Society of Acoustic Technology** is to hold a Conference on "Planning and Design for Protection from Noise" at Coventry on 15 to 18 June 1964.

Further details may be obtained from: The Secretary, The Society of Acoustic Technology, Department of Pure and Applied Physics, The Royal College of Advanced Technology, Salford 5, Lancashire.

**An order of the Privy Council** made on 26 February 1964 now allows amendments to the Charter of The British Institution of Radio Engineers to change its name to The Institution of Electronic and Radio Engineers, so as to more aptly describe the scope of the Institution's work and the professional activities of its members.

**Exports of electronic valves** and semiconductor devices during the last quarter of 1963 amounted to approximately £2 838 000 according to figures issued by The British Radio Valve Manufacturers' Association (BVA), and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA).

The total exports for 1963 were approximately £11¼M, an increase of 4·6 per cent on the corresponding figure for 1962.

**The Bell Telephone Company of** Canada has awarded a contract to the Canadian General Electric Co. Ltd for the supply of equipment for a new microwave system to link Quebec and Riviere-du-Loup with existing Bell Telephone microwave routes. The equipment, which will be manufactured by G.E.C. (Telecommunications) Ltd at Coventry, includes radio, multi-channel i.f. switching, and ancillary items for the system, and will be delivered in the autumn of this year.

The initial system will be 'twin-path', with one working channel and one protection channel, but it will be capable of expansion to its maximum capacity of six working and two protection channels.

The radio equipment will operate in the frequency band 5925 to 6425Mc/s and each radio channel will have a capacity of up to 1800 speech channels. It will meet all the requirements of the C.C.I.R. "circuit fictif" when carrying 1 800 speech channels or a monochrome or colour television programme on each radio channel.

The radio-protection switching equipment is capable of switching one protection-channel for three working channels.

The overall equipment for the system willl be similar to that ordered by the British Post for new microwave links between Southampton, London, Birmingham, Manchester and Carlisle.

**"Measurement of the electrical properties of electronic tubes and valves"** is IEC Publication 151 and the first four parts have recently been issued.

Part 1 deals with the measurement of electrode current; Part 2 with measuring heater or filament current; Part 3 with the measurement of equivalent input and output admittances, and Part 4 with methods of measuring noise factor. The first three parts outline the conditions to be followed in measurement but do not give measuring methods. The methods described in Part 4 are based on current practice.

Copies of all these publications may be obtained from the British Standards Institution, Sales Branch, 2 Park Street, London W.1. (Part 1, 4/6; Part 2, 5/8; Part 3, 8/6; and Part 4, 10/2).

**The Bureau of Naval Weapons of America** is to hold an international Symposium on "Rotating and Static Precision Components" at Washington, D.C. on 21 and 22 April, 1964.

The two-day technical programme is planned to cover the most current and up-to-date information on component technology as well as a projection of this technology to future weapon systems design.

Further information can be obtained from Neal J. Whitney, John I. Thompson & Co., P.O. Box 19006, Washington, D.C.

**The first direct large capacity submarine telephone cable** between Britain and Germay has recently been brought into operation,

The new cable has been laid between Winterton, in Norfolk, and Borkum Island and thence to Leer on the German mainland. It is about 200 nautical miles long and will provide 120 telephone circuits, any one of which will carry 24 telegraph circuits.

The cable together with 20 submerged repeaters, terminal communication and power feeding equipment was manufactured by Standard Telephones and Cables Ltd who were also responsible for the complete cable laying operations which were carried out by HMTS Monarch and the Danish cable ship 'Peter Faber'.

With the inauguration of this new cable subscribers on London STD exchanges will soon be able to dial their own calls to Germany: dialling from other cities in Britain and in the reverse direction will be possible later this year.

To meet the increasing demand for telephone and telegraph services another five new submarine telephone cables will be laid across the North Sea in the next few years. A second Anglo-German cable and a cable between Britain and Denmark will come into service later this year. Two new cables between Britain and the Netherlands will be completed in 1965 and 1966. Another cable is planned between the United Kingdom and Norway.

**The Norwegian Civil Aviation Authority** has ordered for the Fornebu Airport, Oslo, Norway, a radar simulator for the training of new traffic control officers, from the Military Systems and Simulation Division of the Solartron Electronic Group Ltd, Farnborough, Hampshire.

The equipment for Fornebu Airport will train air control officers in their duties. Up to six 'targets' or aircraft at a time will be presented on surveillance radar or p.p.i. displays.

The addition of other units in the future will enable the equipment's scope to be extended for training in precision approach radar.

The simulator enables training to be done as wholly simulated or locked into live airfield radar in the more advanced section of the course.

**Cossor Electronics Ltd** has secured a contract from British European Airways for the supply of the secondary surveillance radar type 1600 transponder. Over a hundred of these equipments are to be installed in the B.E.A. fleet which will include the new Hawker Siddeley Trident jet airliners, due to commence scheduled services this month.

The S.S.R. 1600 transponder complies with the stringent requirements of the B.E.A. specification for transponder equipment.

This airborne transponder also complies fully with the requirements of the A.R.I.N.C. Specification No. 532D. In recent weeks Cossor have also received orders for the 1600 transponder from B.O.A.C. and other airline authorities.

**Russian engineers** in Leningrad have developed an electronic device known as Golos ('Voice') designed to restore human speech. It is of an original and at the same time simple design and weighs only 150g.

It is used mainly by patients who have had to have their vocal chords and a part of the larynx removed. The apparatus generator plays the part of the vocal chords and the sounds are conveyed by a mouthpiece into the hollow of the mouth where they are articulated into coherent speech. On the body of the mouthpiece there is a small timbre regulator which makes it possible for the speaker to impart emotion to his speech.

The new device has already been put into serial production. It will be obtainable through the Leningrad offices of th Chief Medical and Sanitary Equipment Sales Organization.

**Siemens & Halske A.G.** have developed an inductive train control system in conjunction with the West German Federal Railway Administration and a trial installation is now undergoing tests on the line between Hamburg and Forchheim.

The system is designed to control trains travelling at speeds of 200km/h and is necessary on account of the fact that the distance between existing signals is too short for the braking distance of 2500m required for these high speed express trains.

Data regarding distance and aspect of signals up to 6000m ahead is displayed on a panel in the driver's cab and the actual speed of the train is compared with the speed based on the input data. Should the actual speed be excessive the train is automatically braked.

**Decca Radar Ltd** has received a contract from the Ministry of Aviation for the supply and installation of twelve type AR-1 multi-purpose air traffic control radar systems.

Introduced at the end of 1963, the AR-1 is the first surveillance radar designed from the outset to meet all air traffic control surveillance requirements from very close range out to a range of 75 miles. Performance standards in each of its many roles are superior to those demanded by I.C.A.O. international regulations.

The radar systems being supplied to the Royal Air Force will incorporate dual transmitters operating in frequency diversity to provide instantaneous standby in the event of failure. The display units selected are the transistorized autonomous type standard with the AR-1. Each installation will have three units.

The AR-1 radar is a 10cm, 650 kilowatt system incorporating the following technical features:

- Cosecant$^2$ aerial system to give gap-free coverage. Rotation rate 15rev/min.
- Variable polarization (linear/circular) for the suppression of weather returns
- Transistorized m.t.i. system of high stability for the elimination of permanent ground echoes
- Dual diversity transmitters to provide enhanced performance and instant standby in the event of failure
- Rationalized layout and intercabling system to simplify installation at smaller airports.

**The National Physical Laboratory,** is to hold a symposium on the communications, surveying and medical applications of lasers at Teddington on 23 to 24 April.

This meeting will be an informal one, with the first day devoted to applications of gas lasers and the second to those of the solid state. Papers given will be very short (limited to ten minutes), the emphasis being on discussion. There will however be surveys of American as well as British developments.

Applications should be made to the Secretary, Laser Symposium, National Physical Laboratory, Teddington, Middlesex. Attendance will be by invitation only and will be restricted to a maximum of 300.

This symposium is the second in a series of three on laser subjects. The first, dealing with the physics of laser sources, is to be held by the Institute of Physics at Imperial College on 9 to 10 April and the third will be a full-scale symposium at the Institution of Electrical Engineers on 29 to 30 September and 1 October.

# LETTERS TO THE EDITOR

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

## Quantitative Analysis of the Time Response of Flywheel Circuit in Television Receivers

Dear Sir,—In his article D. Maurice[1] analyses the flywheel circuit in a television receiver. For the qualitative analysis of the properties of a flywheel circuit the author uses the oscillograms of the responses for unit step deviation of line-sync frequency. I think, that to measure the properties of the flywheel circuit it would be more advantageous to use the response to unit-step change of input phase since it permits not only a qualitative but also a quantitative analysis of the flywheel circuit. From this response one can determine the noise bandwidth, which directly determines the noise properties of the circuit and damping.

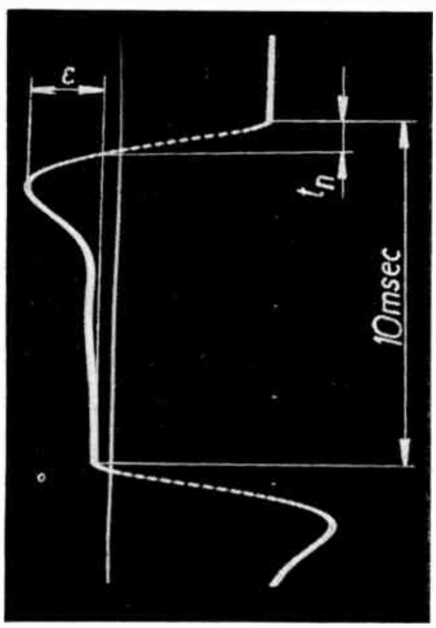
The noise bandwidth is defined by the equation:

$$\Omega_N = \int_{-\infty}^{\infty} |P(j\omega)|^2 d\omega,$$

where $P(j\omega) = \Theta_2(j\omega/\Theta_1(j\omega)$ is the transfer function of the flywheel circuit, $\Theta_1(j\omega)$ is the input and $\Theta_2(j\omega)$ the output phase change.

The theory of derivation of the following graphs for the analysis of the time response is lengthy and has been described in full[2]. Consider an example from which the method of analysis became apparent. Fig. 1 is a photograph of the response of a flywheel circuit to a unit-step change of phase. From the response one can determine the overshoot $\epsilon = 0{\cdot}48$.

Fig. 1. Response of flywheel circuit to unit-step change of input phase



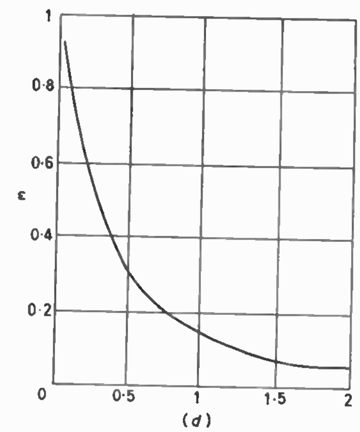




Fig. 2. Overshoot ε of flywheel synchronization as a function of damping d

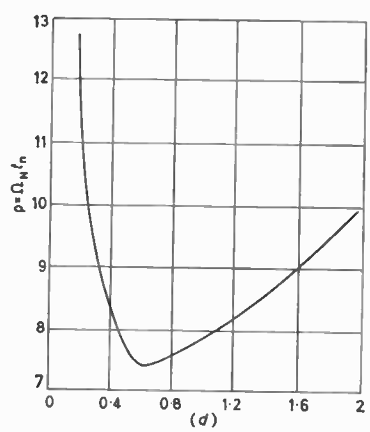



Fig. 3. Graph for determining the noise bandwidth of flywheel synchronization

For this overshoot read off on Fig. 2 the damping of the system $d = 0{\cdot}3$. One can easily determine the rise-time $t_n$ (0 to 100 per cent of the stabilized value) from Fig. 1 in view of the considerable overshoot and the known modulating frequency (50c/s); $t_n = 8{\cdot}55 \,.\, 10^{-4}$sec. For a damping $d = 0{\cdot}3$ we read off $\rho = 9{\cdot}4$ on Fig. 3. Since $\rho = \Omega_N t_m$, the noise bandwidth $\Omega_N = \rho/t_m = 1{\cdot}1 \,.\, 10^4$rad/sec.

Yours faithfully,

Z. Sobotka,

Prague, C.S.S.R.


REFERENCES

1. Maurice, D. Characteristics of Flywheel Synchronizing Circuits in Television Receivers. *Electronic Engng.* 34, 77 (1962).
2. Sobotka, Z. Automatiká fázová synchronisace Automatic Phase Control). Praha, 1963, NCSAV.


### The Author replies:

Dear Sir,—Ing. Z. Sobotka suggests that when analysing the behaviour of a flywheel synchronizing circuit, it is better to apply a unit-step change of phase to the synchronizing signal rather than a unit-step change of frequency.

Mr. Rout and I would still prefer to use a unit-step change of frequency for the following reasons. If the circuit under examination behaves in a linear fashion when the reference synchronizing signal is perturbed by small phase or frequency deviations, then the response to a unit-step change of phase can be obtained by differentiating, with respect to time, the response to a unit-step change of frequency. However, in order to obtain the response to a unit-step change of frequency, given the response to a unit-step change of phase, it is necessary to integrate the response and the constant of integration, which is the loop-gain of the system ($\beta_1\beta_2$rad/msec per rad in our article), cannot readily be determined. The accurate determination of loop-gain is particularly important when the response of flywheel synchronizing circuits to low frequency disturbances (e.g. mains-frequency modulation of synchronizing-pulse timing) is being considered.

Yours faithfully,

R. D. A. Maurice,

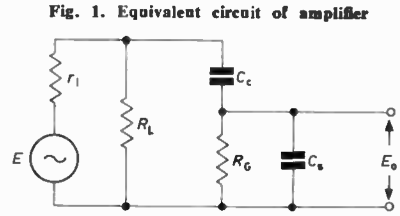
British Broadcasting Corpn., Tadworth, Surrey.

## A Limitation of Current-Feedback

Dear Sir,—In many amplifier stages it is common practice to use current negative feedback. While the application of current-feedback has many beneficial effects it may have the undesired effect of reducing the high frequency response of the amplifier if the load impedance is capacitive. An examination of several textbooks shows that this aspect of current negative feedback has not received attention. On the contrary, stress is laid on the improvement of the high frequency response by the use of negative feedback, the inference being that it applies to both the current and the voltage feedback cases.

The equivalent circuit of an amplifier is shown in Fig. 1, the internal generator

Fig. 1. Equivalent circuit of amplifier



being represented by a voltage $E$ and a resistance $r_i$. The output voltage $E_o$ at frequencies which are sufficiently high for the reactance of the coupling capacitor $C_o$ to be negligible, is

$$E_o = \frac{Eg_i}{g_i + G_L + G_G + j\omega C_S} \quad \dots\dots\dots \text{(1)}$$

The upper frequency $f_\mu$ at which the voltage amplification is reduced by 3dB is

$$f_\mu \frac{g_i + G_L + G_G}{2\pi C_S} \quad \dots\dots\dots \text{(2)}$$

Rearranging equation (2) to show the effective time-constants $C_S r_i$, $C_S R_L$, $C_S R_G$

$$f_\mu = 1/2\pi(1/C_S r_i + 1/C_S R_L + 1/C_S R_G) \quad \dots\dots\dots \text{(3)}$$

Thus in the case of current feedback where the internal resistance $r_i$ is increased we see that this increases the time constant $C_S r_i$ and reduces the upper frequency response. This of course is the opposite result to that obtained using voltage feedback where $r_i$ is reduced and the high frequency response is increased.

Conversely however it will be seen that at low frequencies, where the reactance of $C_S$ is sufficiently high to be neglected, the low frequency response will be improved by the application of negative current feedback. The corresponding equation to those of (2) and (3) is

$$f_L = \frac{1}{2\pi C_C}\left[\frac{1}{r_i R_L/(r_i+R_L)+R_G}\right]$$

$$= 1/2\pi\left[\frac{1}{C_C(r_i R_L/(r_i+R_L))+C_C R_G}\right]$$

Here the use of voltage feedback would degrade the low frequency response though this can be overcome by increasing the value of the coupling capacitor.

It is assumed here that the feedback loop does not involve any additional phase shift.

Yours faithfully,

G. May,

Royal Military College of Science,
Shrivenham.

## Bridge Rectifier Symbols

Dear Sir,—May we suggest to your readers a graphical symbol for a bridge rectifier which, while not having the merit of following standard practice, seems to us to save a considerable amount of trouble in drawing and to facilitate a somewhat neater presentation of circuit diagrams.

The commonly used symbol shown in Fig. 1(a) is awkward to draw and places the a.c. and d.c. lines in an inconvenient relationship; the latter disadvantage is not eliminated in the simpler version shown at (b), although it is less troublesome to draw. A different arrangement, shown at (c), has the big advantage of having the input and output circuits into

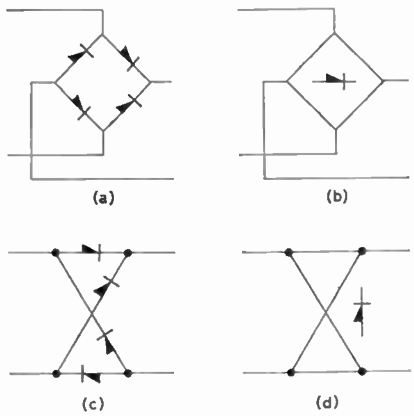



Fig. 1. Bridge rectifier symbols

line, but is if anything more tedious to draw than (a).

The proposed symbol is that shown at (d). Input and output are conveniently in line, and the single diode symbol shows clearly which is the d.c. side and which way the rectifier is poled, and strongly suggests the free-wheeling path.

Yours faithfully,

K. G. King,

R. H. Eastop,

Electrical Research Laboratories,
Westinghouse Brake & Signal Co. Ltd.,
Radlett, Herts.

## Computation of Inverse Impedance by De Morgan's Theorem

Dear Sir,—De Morgan's theorem is extensively used to find the complement of a Boolean expression. The theorem can be stated as follows:

$$\overline{A + B} = \overline{A}\ .\ \overline{B}$$

$$\text{and} \quad \overline{A\ .\ B} = \overline{A} + \overline{B}$$

where $\overline{A}$ and $\overline{B}$ are complements of $A$ and $B$ respectively.

'+' means AND, so that $A+B$ means $A$ AND $B$, and . means OR so that . $A\,.\,B$ means $A$ OR $B$. Many times $A\,.\,B$ is simply written as $AB$.

The theorem can be very conveniently used to find the inverse impedance of a two terminal network composed of several elements. To do this one should be able to write the network as a Boolean expression. For this the series connexion is indicated by + and parallel connexion by . . Thus the network of Fig. 1 can be written as follows:

$$Z = A+(B+C)D \quad \dots\dots\dots \text{(1)}$$

Its inverse is the Boolean complement of $Z$, so that,

$$\overline{Z} = \overline{A+(B+C)D} \quad \dots\dots\dots \text{(2)}$$

The right-hand side can be simplified by De Morgan's theorem, yielding,

$$\overline{Z} = \overline{A}\,.(\overline{B}\,.\overline{C}+\overline{D}) \quad \dots\dots\dots \text{(3)}$$

The Boolean complement of each individual element should be computed beforehand from the usual relations as given in Table 1.

TABLE 1

| Element | Inverse with respect to a constant $K$ |
|---|---|
| Resistance, $R$ | $R' = K^2/R$ |
| Inductance, $L$ | $C' = L/K^2$ |
| Capacitance, $C$ | $L' = K^2C$ |

Thus the inverse impedance of Fig. 1 is shown in Fig. 2. This is obtained by

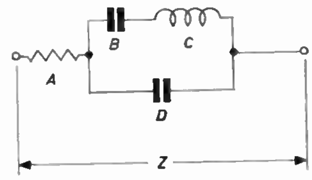



Fig. 1. Two terminal network

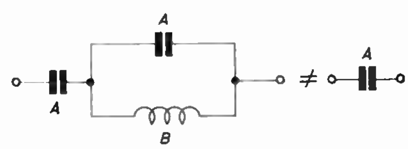



Fig. 2. The inverse of Fig. 1

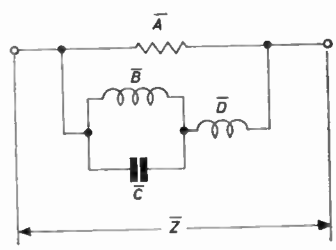


Fig. 3. An electrical network

writing equation (3) in the circuit form, remembering that + means series connection and . parallel. It is true that the circuit of Fig. 2 can be drawn by applying the rule of converting a series connexion into a parallel one and vice versa. In fact, this is exactly what the De Morgan's theorem does. But the fact that an electrical network can be expressed as a Boolean equation has its obvious advantages, especially when the network consists of several elements.

It should be mentioned here that no attempt is to be made to simplify the Boolean expression of the network before and after the application of De Morgan's theorem. This is because the simplifying theorems do not hold good in case of electrical networks, as is illustrated in Fig. 3, wherein it is shown that the theorem $A+AB=A$ of Boolean Algebra is not true in case of the network.

Yours faithfully,

N. N. Biswas,

Dept. of Electrical Engng.,
University of Roorkee,
Roorkee, U.P., India.

# ELECTRONIC EQUIPMENT

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

(Voir page 275 pour la traduction en français; Deutsche Übersetzung Seite 282)

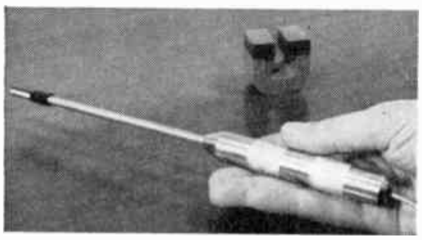
### HALL PROBE KIT

**Scientifica, 148 St. Dunston's Avenue, Acton, London, W.3**

*(Illustrated below)*

By utilizing the Hall effect, this kit provides a simple and convenient means of measuring magnetic field strengths. By connecting the output of the Hall probe direct to a 50$\mu$A meter, field strengths from 100 to 20 000 gauss can be easily measured. The lower end of this range can however be further extended by use of a single stage transistor amplifier for which a circuit is detailed in the instruction manual.

The Hall probe is supplied with a fully illustrated manual which contains easy to follow diagrams illustrating the simple steps to follow in assembling the probe, and also a clear and thorough description of the theoretical principles involved. This will enable students to



conduct numerous experiments involving the Hall effect, and at the same time appreciate the practical significance of this discovery.

Calibration of the Hall probe is provided by a standard magnet supplied with the kit. Alternatively the standard magnet may be used in conjunction with the probe to demonstrate the Hall effect, or for measurement of the Hall parameters.

The output for a 10kG field is 200mV and the temperature stability is 0·3 per cent/°C.

**EE 68 751** *for further details*

### SNAP-ACTION DELAY SWITCH

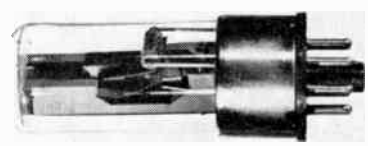
**Standard Telephones & Cables Ltd., Connaught House, 63 Aldwych, London, W.C.2**

*(Illustrated above right)*

A new snap action thermal delay switch, known as the S45 C/ID is now available from Standard Telephones and Cables Ltd. Designed for switching comparatively high powers (up to 250V at 5A, a.c.), the new switch is a single-pole changeover device featuring high contact pressure and a rapid make and break action.

Nominal switching delay time is 42sec, and by connecting a number of switches in series, multiples of this delay time can be obtained.

Thermal delay switches are very suitable for providing the delay between the



application of heater voltage and other circuit voltages, such as anode voltage, to indirectly heated valves and tubes. In addition they can be used as v.l.f. relaxation oscillators, for switching three-phase circuits from star to delta arrangements when starting induction motors and for re-closing a circuit-breaker after a temporary current surge has caused it to trip.

**EE 68 752** *for further details*

### PRECISION INTERVAL TIMER

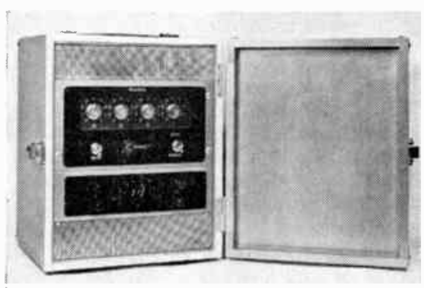
**Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex**

*(Illustrated below)*

This interval timer has been designed specifically to measure in milliseconds the operating times of relay contacts but it can be used equally well for the determination of any time interval delineated by an external contact or contacts.

Internal pulses at a rate of 1 000 per second derived from a crystal-controlled oscillator (accuracy 0·01 per cent) are counted by four Dekatron tubes in cascade during the period that the external contact(s) are operated and the resultant time (up to a maximum of 9 999sec) is read directly from these Dekatron tubes.

A six-position selector switch and two pairs of spring-loaded input terminals allow the timer to be started and stopped by any combination of one or two contacts either opening or closing. A three-position 'reset-operate' switch is fitted



## HANOVER FAIR

**'Electronic Engineering' will occupy Stand 2 in Hall 13 at the Hanover Fair from 26 April to 5 May and visitors will be welcome.**

and the equipment runs from 200 to 250V a.c. mains supplies.

Several alternative versions are available with maximum counts up to 999 99sec (in hundredths of a second) and all types can be supplied with or without a polished wooden carrying case.

Provision can also be made for operation by external pulses or d.c. voltages at a small extra charge.

**EE 68 753** *for further details*

### BRIDGE 'AUTOBALANCE'

**The Wayne Kerr Laboratories Ltd, New Malden, Surrey**

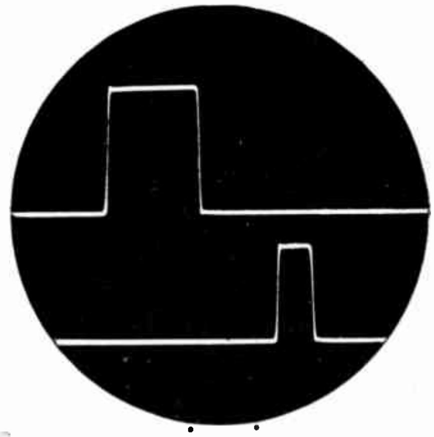
*(Illustrated below)*

A new instrument developed by The Wayne Kerr Laboratories Ltd enables accurate measurements of the resistive

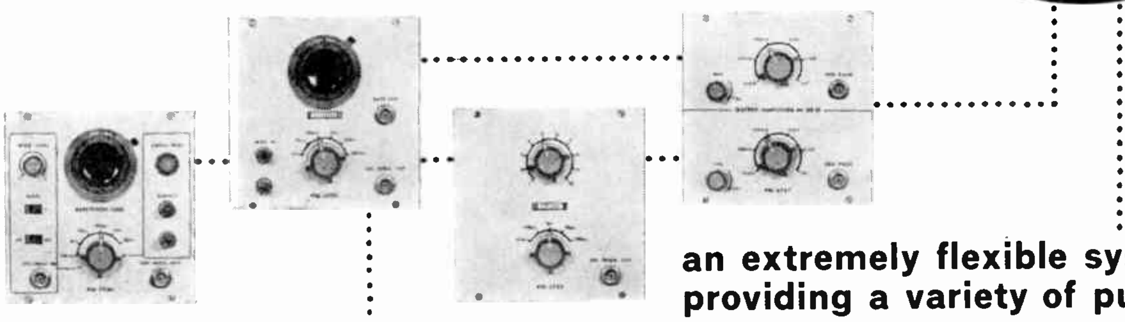
and reactive terms of an unknown impedance to be made without manual balancing. Operated in conjunction with the Wayne Kerr universal bridge type B221, the new 'Autobalance' adaptor AA221 electronically maintains a constant state of balance, presenting variations in the value of either term immediately on a twin-meter display.

The two meters on the Autobalance adaptor effectively replace the vernier controls on the bridge. Thus, by suitable choice of bridge range, the meters can provide either the entire reading or, with the first two digits set-up on the bridge decade switches, an overall measurement to four significant figures. In either case the vernier controls are set to zero and the final balance maintained electronically. The basic 0·1 per cent accuracy, exceptional range and discrimination of the B221 bridge are unaltered when it is used with the Autobalance adaptor.

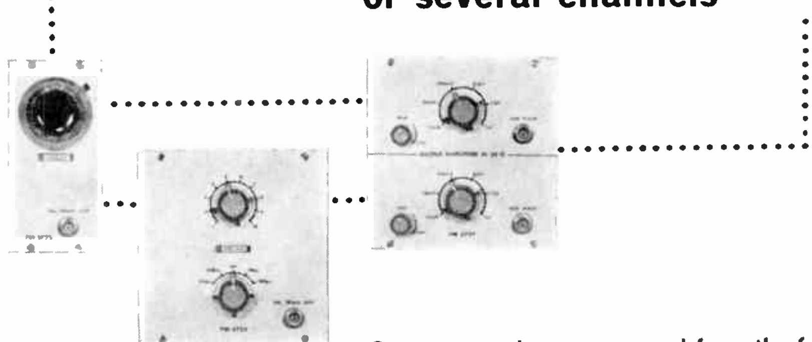
For batch testing, the bridge vernier controls can be so adjusted that the meters read mid-scale at the desired nominal value. By using a simple mask on the meter scales, components can be checked using unskilled labour. Outputs from the metering circuits are also brought out to jack sockets at the rear of the Autobalance adaptor for the



# new modular pulse generator with unique features



**an extremely flexible system, providing a variety of pulses and pulse sequences on one or several channels**



**Systems can be programmed from the following units:**

**Master generator PM 5720:** 10 c/s – 10 Mc/s
External triggering DC-15 Mc/s; single shot/remote control

**Short delay/width unit PM 5722:** 10 ns – 1 ms

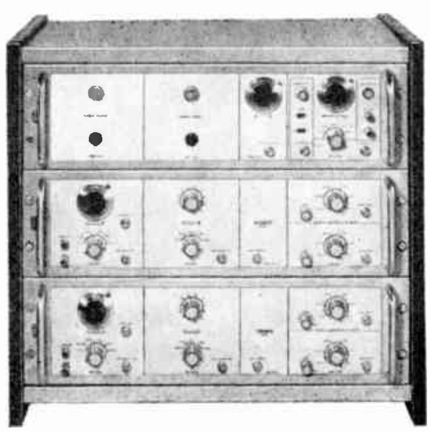
**Long delay/width unit PM 5723:** 1 μs – 1 s
External gate output; facilities for modulation of pulse delay/width

**Interpulse unit PM 5725:** For additional setting of pulse delay/width in connection with PM 5723

**AND-OR gate PM 5732:** OR and AND-gate for system needle pulses

**Output unit PM 5727:** Output 5 mV – 5 V across 50 Ω
Rise and fall time < 10 ns; true positive and negative pulses simultaneously available

**Power supply unit PM 5740:** 110 – 245 V, 50 – 60 c/s



PHILIPS

# PHILIPS

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operation of automatic batching equipment. It is also possible to operate digital voltmeters and print-out mechanisms, recorders and remote-reading meters.

The Autobalance adaptor provides both signal source and detector, transistorized and powered by an internal battery. The combination of bridge and adaptor can therefore be used where a.c. power is not available. For continuous use, a mains-operated power supply can be supplied.

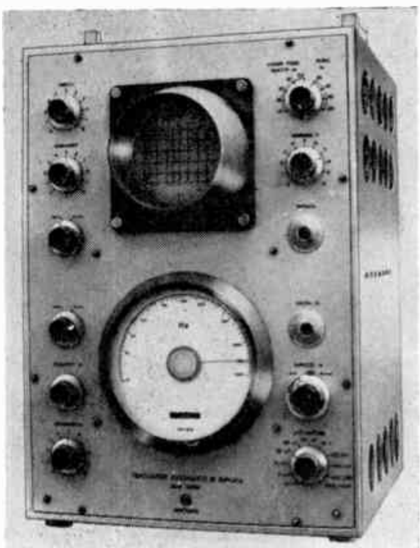
**EE 68 754** *for further details*

## A. F. FREQUENCY RESPONSE TRACER

**Distributed by: Claude Lyons Ltd, Hoddesdon, Hertfordshire**

*(Illustrated below)*

Manufactured by Construzione Elettriche di D. Borsini of Ancona, Italy, the TAR/61 automatic frequency response tracer is an audio frequency spectrum analyser which simplifies experimental



work on, and testing of amplifiers, filters, recorders, microphones, pick-ups, loudspeakers and transducers, and circuit elements generally.

The instrument essentially comprises a low distortion beat frequency oscillator, covering the range 0 to 20 000c/s, which is swept repetitively by a motor drive. The variable frequency, constant level output of the b.f.o. is applied, via a power amplifier with calibrated attenuators, to the device under test, and the output of the latter is returned to the tracer where it is applied to a logarithmic amplifier feeding the vertical deflexion system of a long-persistence cathode-ray tube. The horizontal sweep for the c.r.t. is derived from a very robust 180-step potentiometer mounted on the b.f.o. sweep drive shaft, thus producing a direct amplitude/frequency graph.

The oscilloscope graticule is calibrated vertically in decibels and millivolts, and horizontally in frequency, and the trace persistence is sufficiently long to permit successive sweeps to be superimposed for direct comparison, thus making immediately evident the result of any adjustment. Reference response curves, or calibration limits, can be marked on the graticule, thus greatly simplifying production testing.

Where two-terminal devices such as microphones or loudspeakers are to be examined, it is of course necessary to use a corresponding transducer to complete the system—a loudspeaker to produce the energization for the microphone for instance. The secondary transducer may produce errors in the results, and provision is made for the connexion of a compression amplifier in the oscillator circuit, fed from an auxiliary transducer, to correct for any non-linearity.

**EE 68 755** *for further details*

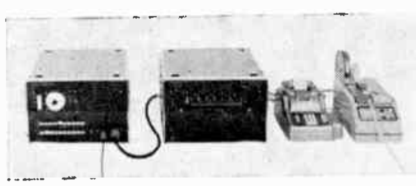
## STRAIN AND TEMPERATURE RECORDER

**Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex**

*(Illustrated below)*

Accommodated in two units, this transportable data logging system has been designed under contract to the D.S.I.R. and provides facilities for strain and temperature measurement in prestressed concrete structures.

Up to fifty channels of each parameter can be measured, and the output is presented digitally both as a printed record and on punched tape.



Strain is measured on the vibrating wire strain gauge. This gauge is designed to be embedded in the concrete structure. The period of 100 cycles of wire vibration is recorded and the simple square law relationship between strain and period thus measured, facilitates direct calculations of the internal strain in the structure.

Temperature measurements are derived from resistance thermometer elements and the results are displayed as direct readings in the absolute scale of temperature.

The equipment is designed for unattended site operation and can be programmed to run automatically at predetermined intervals of up to 2 hours.

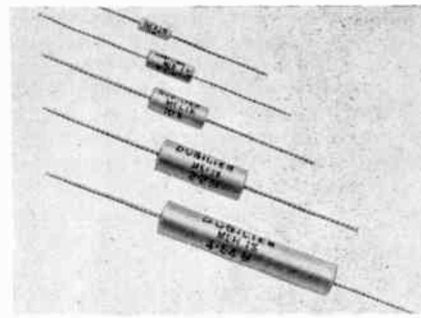
**EE 68 756** *for further details*

## METAL FILM RESISTORS

**Dubilier Condenser Co. (1925) Ltd, Ducon Works, Victoria Road, North Acton, London, W.3**

*(Illustrated above right)*

Type ME metal film resistors are available in five wattage ratings and resistance values of 30Ω to 10MΩ subject to range limitations per wattage size. The standard tolerance is ±1 per cent but closer tolerances of 0·5 per cent, 0·25 per cent and 0·1 per cent are available



dependent upon temperature coefficient.

The outer casing for the lower wattage sizes is a moulding of special moisture resistant high temperature plastic insulation and for the 1 and 2W ratings a porcelain tube with hermetic end seals. All five sizes have very low noise level independent of the resistance value, normally below 0·1$\mu$V/V.

The voltage coefficient when measured between one-tenth and full rated voltage is less than $5 \cdot 10^{-6}$/V for all values. On low and medium value resistors the voltage coefficient is essentially zero. Data indicates that changes over a one-year period of shelf life at normal room temperature cannot be measured on standard laboratory equipment. They can be supplied in five different temperature coefficients from $\pm 150 \cdot 10^{-6}$ down to the close efficient of $\pm 15 \cdot 10^{-6}$/°C.

The resistors are non-inductive up to approximately 100Mc/s except in the case of the low value high wattage types. They are very suitable replacements for precision wire wound types where miniaturization is of importance.

**EE 68 757** *for further details*

## BEAT FREQUENCY OSCILLATOR

**Distributed by: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1**

*(Illustrated below)*

An oscillator covering the entire audio and sub-audio frequency bands in a single range is available from Livingston Laboratories Ltd. It is the type HO32 beat frequency oscillator manufactured by Radiometer of Denmark.

The instrument has an overall frequency range of zero to 21kc/s, the range from zero to 20·5kc/s being covered in a single sweep. The maximum output power is 4W which is available from an output transformer or from the calibrated attenuator. The transformer has tappings for impedance matching of various loads with the option of a 15Ω

output. The attenuator provides output levels from 300$\mu$V to 100V. The harmonic distortion is less than 0·1 per cent for attenuator outputs below 10V. Good frequency stability and response, and a magic eye indicator for easy determination of zero beat are further features of this instrument.

**EE 68 758** *for further details*

### INSULATION METER

**Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex**

*(Illustrated below)*

The type 190 insulation meter is designed for the direct measurement of insulation resistance and leakage current. The instrument is portable and uses a transistor invertor to generate the test voltages from an internal battery. Test voltages of 500V, 250V and 100V are provided. The use of a meter with a 3½in scale length allows insulation resistance of 1000MΩ to be measured.

Push-button on-off switching gives instant one-hand operation and eliminates unnecessary battery drain. A single rotary switch is used to select the test voltage required and includes a 'calibrate' position for setting the output voltage with a preset control. The use of a mercury battery as a compact source of power gives a life expectancy of more than 25 000 average operations with long shelf life and freedom from corrosion.

The output is electronically stabilized and the maximum current is limited to 50$\mu$A. This current limiting circuit reduces the test voltage proportionately from its full value under open-circuit conditions to zero at full scale. This feature virtually eliminates the risk of destructive breakdown and allows the instrument to be safely used for the measurement of leakage current of semiconductor rectifiers, diodes etc.

**EE 68 759** *for further details*

### PRESSURE TRANSDUCERS

**Intersonde Ltd, The Forum, High Street, Edgware, Middlesex**

*(Illustrated above right)*

To measure static or dynamic gaseous or liquid pressure Intersonde Limited


has introduced the types PR 14 and PR15 bonded strain gauge pressure transducers to cover pressure ranges extending from 0 to 1 000lb/in² to 0 to 10 000lb/in².

The pressure responsive element common to both types is an accurately machined tube on which bifilar strain gauge elements are wound and connected in a four-arm bridge configuration. Both transducers have an output resistance of 350Ω and produce an output of 22mV at full rated pressure with 15V excitation. The combined non-linearity and hysteresis error is within 0·25 per cent of full range and the operating temperature limits extend from −40°C to +120°C.

The type PR14 is provided with pin terminals and the type PR15 with an integral flying lead, both transducers measure 4·9in long by 1·06in diameter and both have a 1/4in B.S.P. pressure inlet.

**EE 68 760** *for further details*

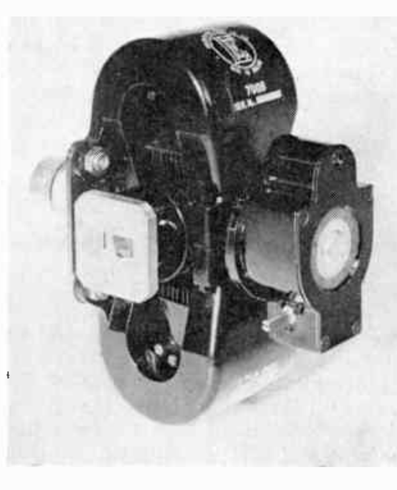
### TUNABLE PULSE MAGNETRON

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

*(Illustrated below)*

The 7008 is a tunable magnetron with electrical characteristics similar to those of the 4J50A series. It has a typical peak output power of 220kW and a frequency range of 8500 to 9600Mc/s. It has a duty cycle of ·001 with a pulse width of 0·2 to 2·75$\mu$sec at a peak anode current of 27·5A.

Special features of the 7008 include:




minimum peak output power over whole of frequency range is 200kW; linear tuning characteristic with coarse and fine calibrated dials; the compact tuning mechanism is specially designed to have small backlash; no moding; missed pulse rate less than 0·2 per cent; high rate of rise of voltage; special long life cathode; little or no sparking; unique heater construction for greater reliability; cooling by natural convection under certain environmental and operational conditions.

**EE 68 761** *for further details*

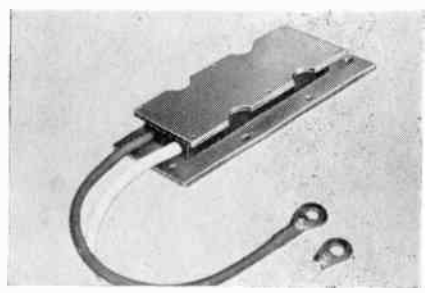
### 'PELTIER BATTERIES'

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

*(Illustrated below)*

Mullard Ltd has introduced three 'peltier batteries' suitable for a wide range of cooling applications where a compact 'cold source' is required.

In situations where space is limited, the Peltier battery offers considerable advantages over conventional refrigeration apparatus since it requires neither



a compressor nor a heating unit. Typical uses are in cooling and small scale refrigeration in medical and biological research and heat sinks in transistor equipment.

The batteries which use bismuth telluride as the thermoelectric element, operate on the Peltier principle that when a direct current is passed through a junction of dissimilar metals, or semiconductor materials, a temperature gradient is established across that junction.

Battery type PT11/20 is designed for an operating current of 18 to 22A at 1·0 to 1·2V, and has a maximum cooling capacity of 16W. Its minimum life expectancy is 2 000 hours of continuous operation at 20A.

Type PT20/20 has an operating current of 20A at 2V, a cooling capacity of 23W, and a minimum life expectancy of 2 000 hours of continuous operation at 20A.

Type PT47/5 operates at 5 to 6A and 5·0 to 5·4V and has a cooling capacity of 16W. Its minimum life expectancy is 2 000 hours of continuous operation at 5A.

Both the PT11/20 and PT47/5 are available with a flat copper plate for use with solid surfaces, or with fins for the cooling of gases or liquids. Type PT20/20 has flat copper plates.

All three types are supplied ready for immediate use.

**EE 68 762** *for further details*

**Leadership in Semiconductors**

**TEXAS INSTRUMENTS**

# Can you afford transistor failure?

## Some plain facts about reliability

A six-transistor radio using transistors with a failure rate of one device per 260,000 hours should play five years before failure. But given the same failure rate, a 26-transistor television set would have a predicted trouble-free life of a year and a half. And a 40,000-transistor computer might remain trouble-free for only seven hours. (A similar valve computer might require attention every hour.)

Modern semiconductor technology has made it possible to meet the requirements of the computer. But is the problem of reliability now finally a simple one solved equally well by all manufacturers? Has the industry reached the ultimate in reliability assurance techniques?

### Innovation necessary

We believe the answer to both questions is no. Product reliability will continue to depend upon experience, proper execution and innovation.

TI has been the acknowledged industry leader in advancing semiconductor reliability. technology Where equipment or techniques did not exist to do a specific job, TI developed them. The Transistor and Component Tester (TACT) machine shown below is just one example of TI innovation. TACT tests as many as 20 parameters on 7,200 units per hour.

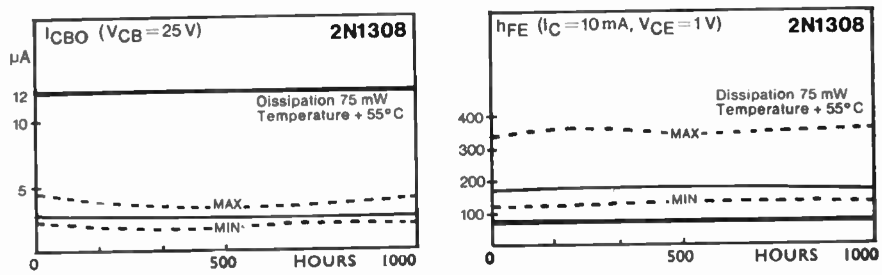
## Progress report on semiconductor networks

SOLID CIRCUIT semiconductor networks, announced by TI in 1958, are moving rapidly toward the widespread use predicted for the next few years.

Today, SOLID CIRCUIT semiconductor networks from TI are used in more than 100 military and industrial equipment programmes around the world to improve reliability while cutting weight and size. Production at TI in 1963 increased eight-fold. TI produced and shipped more integrated circuits in the fourth quarter of 1963 than the entire industry in the second.

And a new catalogue series of high-speed digital networks is now in production .It will be discussed here in detail soon.

## 150,000,000 germanium alloy transistors used in industrial applications alone prove this reliability story...




**Variation of $I_{CBO}$ and $h_{FE}$ over 1000 hours**

These test results for series 2N1302 transistors, taken over 1000 hours, illustrate one part of TI's industry-leading reliability programme. Further proof of TI reliability comes from customer usage records. More than 150,000,000 germanium alloy devices have been manufactured and delivered for industrial applications alone.

Extended life testing of this class of device has shown a demonstrated reliability of less than 1.5 failures for every million transistor operating hours at maximum dissipation stress conditions. Special reliability programmes developed on request have produced even higher standards.

### Applications

Consider these recent applications from TI France where many millions of germanium alloy transistors have been manufactured since 1961.

- Failure in the transistorized process control equipment linked to a strip rolling mill means a £750 hourly loss to a European steel company. Germanium devices with a $10^{-8}$ failure rate history were used.
- A computer manufacturer requested a $10^{-8}$ failure rate with confidence of 90% to guarantee effective performance of a new 10,000 alloy transistor, 20,000 diode model. A comprehensive reliability study by TI produced 2N1308 and 2N1309 germanium devices to meet these requirements.
- When a $10^{-8}$ failure rate objective was asked for high voltage relay driver applications, TI filled the requirement with high-voltage, high-current 2N1924 transistors. Additional benefits: ready availability of the germanium transistors and ease of application.

### Worldwide military acceptance

Many TI germanium transistors including the versatile 2N1302-2N1309 series have earned military approval in the USA, France and Great Britain. The Nice plant of TI France is approved by the Ministry of Aviation to give AID, ARB and related releases and to supply Type Approved (CV prefix) transistors. Similar approval has been given by the French Comité de Coordination des Télécommunications. Here are the corresponding type numbers for the 2N1302 series:

| Commercial Type | British Type Approved | U.S. MIL Approved* |
|---|---|---|
| 2N1302 (NPN) | CV 7348 | USN 2N1302 |
| 2N1303 (PNP) | CV 7352 | USN 2N1303 |
| 2N1304 (NPN) | CV 7349 | USN 2N1304 |
| 2N1305 (PNP) | CV 7353 | USN 2N1305 |
| 2N1306 (NPN) | CV 7350 | USN 2N1306 |
| 2N1307 (PNP) | CV 7354 | USN 2N1307 |
| 2N1308 (NPN) | CV 7351 | USN 2N1308 |
| 2N1309 (PNP) | CV 7355 | USN 2N1309 |

*(MIL-S-19500/126A)

### Why TI?

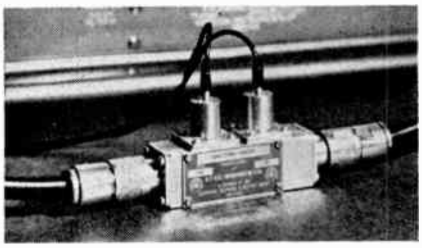
Germanium transistors are widely used and widely available. Specifications and prices are often the same. Why pick TI as your source? Proved reliability is one reason. Confidence that TI's broad line (germanium and silicon) offers you the best transistor for your specific needs is another. For more information on continuing technological and product developments from TI, ask to receive regularly the TI NEWSLETTER.

**TEXAS INSTRUMENTS LIMITED**

MANTON LANE · BEDFORD ENGLAND

TELEPHONE: BEDFORD 67466 · CABLES: TEXINLIM BEDFORD TELEX: 82178

EE 68 144 for further details



## COAXIAL REFLECTOMETERS

**A. T. & E. (Bridgnorth) Ltd, Bridgnorth, Shropshire**

*(Illustrated above)*

A new range of coaxial reflectometers suitable for permanent installation in coaxial lines for monitoring incident and reflected power for users of u.h.f. equipment, has been introduced by A T & E (Bridgnorth) Limited, a Plessey Group Company.

Features include a high power handling capacity and low insertion loss. Back-to-back resistive loop directional couplers make slotted line measurements unnecessary.

Available types cover frequency ranges from 50 to 500Mc/s. The one designed for use in the 150 to 300Mc/s range has a minimum directivity of 40dB. Coupler sensitivity is 500μA for 10W incident power. Maximum primary line standing wave ratio is better than 1·05:1 and there is negligible primary line insertion loss. The power handling capacity is up to 300W.

**EE 68 763** *for further details*

## PRECISION POTENTIOMETERS

**Salford Electrical Instruments Ltd, Peel Works, Salford, 3, Lancashire**

*(Illustrated below)*

Salford Electrical Instruments Ltd have extended their range of precision potentiometers and now include frame size 11. All dimensions including location spigots are to the requirements of this International size.

Resistance values from 500Ω to 50Ω can be supplied, wound on precision machined toroidal formers using nickel-chromium resistance wire or precious metal wire for low noise applications.

The temperature range is −40°C to +85°C and the rating is 2W at 20°C. Independent linearity tolerance is ±0·5 per cent.

The potentiometer is available in ganged form up to four gangs. Extra taps can be fitted, each one being connected to a single turn of wire. Gold plated side terminals are standard for ease of soldering external connexions.

Precious metal dual wiping contacts are used on both the resistance track and the rhodium-plated slip-ring.

The stainless steel spindle is mounted in high quality shielded ballraces.

High quality corrosion resistant aluminium alloy is machined to form the 1·06in diameter case.

**EE 68 764** *for further details*

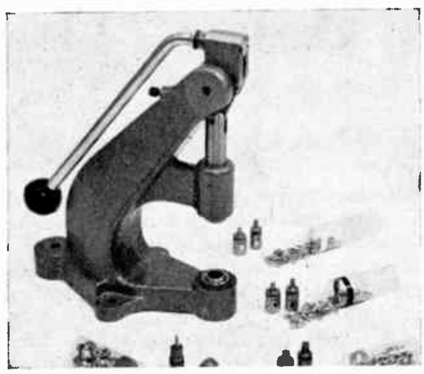
## 'PRESTINCERT' TOOL KIT

**Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex**

*(Illustrated below)*

Belling & Lee Ltd are marketing a new Prestincert tool kit for applications in electronics, electrical work and in light mechanical engineering.

The Prestincert range of components



includes self-punching threaded bushes and terminal posts for single operation mounting on to sheet metal and many synthetic insulation materials. No drilling or riveting is required, and the fixture is virtually indestructible in normal use.

The new kit is designed to meet the growing demands of design engineers and technicians, who wish to make use of workshop techniques during development and to include time saving Prestincert components in their plans for future production lines.

Comprising a small, manually operated, lightweight press—developing up to 0·5ton/in²—the Prestincert tool kit includes six containers of Presincert components of the sizes most commonly used in the light engineering industries.

There are three sizes of threaded bushes—6 B.A, 4 B.A. 2 B.A.— and three types of terminal posts for electrical and electronic applications. Additional supplies of all components are readily available.

Six pairs of insertion tools, colour coded to the appropriate Prestincert components, are also included in the kit. These may be simple push-fitted into the press, and are retained by an integral spring clip.

The tools, press and Prestincert components are packed in a robust transit case.

**EE 68 765** *for further details*

## MAGNETIC RECORDING HEADS

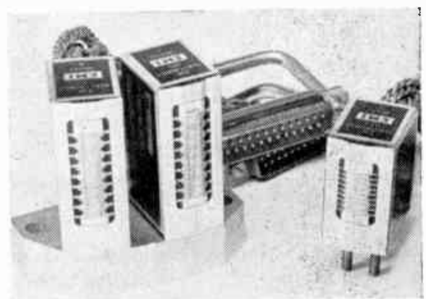
**EMI Electronics Ltd, Hayes, Middlesex**

*(Illustrated below)*

A new range of precision made magnetic heads—the Emidata 'E' series—is now available from EMI Electronics Ltd. These heads have been designed specifically for multi-track instrumentation and data processing systems where high consistency of performance is required.

The exceptional performance characteristics and long life of all heads in a stack are achieved by rigid control of gap depth during manufacture. The heads also have higher order of accuracy of gap alignment and azimuth than the previous 'S' range of heads with which they are completely interchangeable.

Mean azimuth and gap scatter is such that the in-line edges of all gaps in an



assembly lie within a zone 200μin wide and square to the base. The elevation, or head stack tilt, is held within a similar zone 500μin wide. The standard gap lengths are 0·0005in for recording and 0·0002in for playback.

The heads give a high frequency response up to 100kc/s and have a long wavelength response of 0·1in. Crosstalk rejection is extremely good. The record heads write tracks which are wider than the playback heads, allowing for possible tape weave.

All I.R.I.G. and S.B.A.C. track configurations are available up to 33 tracks interlaced on 1in tape. Special track pitch and width can be supplied to special order. The heads are supplied wired to miniature Cannon plugs.

**EE 68 766** *for further details*

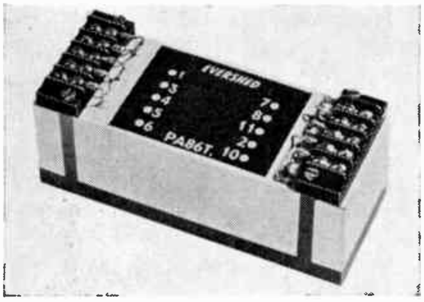
## MINIATURE SERVO AMPLIFIER

**Evershed & Vignoles Ltd, Acton Lane, Chiswick, London, W.4**

*(Illustrated on page 272)*

The Evershed miniature transistor amplifier type PA.86T is designed to drive split-phase a.c. servo motors at frequencies, from 50 to 400c/s, giving a maximum output of 10W at 20-0-20V into an impedance of 35Ω per winding.

The amplifier uses five transistors, the components being mounted within the



heat sink of the output transistors. Complete encapsulation in silicone rubber is utilized as a means of excluding dirt and moisture, and to enable the amplifier to withstand considerable mechanical shock. All connexions are brought out to two screw-connexion strips on top of the amplifier.

For use on an a.c. power supply the amplifier should be operated with an Evershed input transformer type PA.43597, which is rated at 23-0-23V output and can supply power to three amplifiers. Additional windings are provided for motor and generator reference phases and for the chopper transformer (if used).

**EE 68 767** *for further details*

## 'UNSOLDERING' TOOL

**Charles Austen Pumps Ltd, Petersham Works, High Road, Byfleet, Surrey**

*(Illustrated below)*

The unsoldering and removal of faulty components in printed circuits can present difficulties, especially if they are of the multi-pin type. When a large iron is used the adjacent components are affected by heat which can damage the unit beyond economical repair. Short-circuits are sometimes made by excess solder which has become displaced.

In view of the increasing use of printed circuits Charles Austen Pumps Ltd have developed the Austen 'Soldermaster' specifically to overcome these difficulties. This instrument is a small high temperature iron with a hollow bit through which a vacuum may be applied, at the control of the operator, by means of a pump. The solder joint which is melted by the iron is drawn through the bit and trapped in a catchpot. The undesirable flux fumes are also extracted from the vicinity of the joint. The pump is protected by a filter with a porous plastic element which is cleanable. The solidified solder may be emptied from the catchpot at the completion of the operation.

The unit consists of an Austen 'Capex' pump with an Austen Micro-filter, solder catchpot and pipes, a 50W soldering iron with hollow bit and one spare bit. It is supplied complete mounted on a base with an iron rest and carrying handle. Size 7¼in × 6in × 6in high, weight 6lb complete.

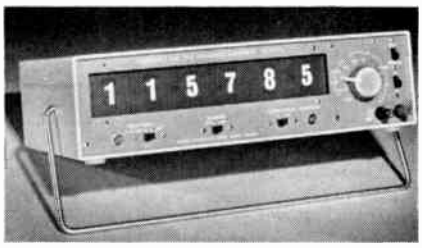
**EE 68 768** *for further details*

## TIME-FREQUENCY METER

**Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey**

*(Illustrated below)*

Venner Electronics Ltd has introduced the TSA 3436, a compact general purpose instrument giving six-digit indication of frequency and period up to 1·2Mc/s, and providing time measurement from 1μsec to 100 000sec. It can also be used as a counter at speeds up to 1Mp/s.



Designed for laboratory use, or for quality control applications in industry, the fully transistorized, portable TSA 3436 measures only 14¼in × 3¾in × 10½in. Main features include three gating times, optional blanking between results, and self-checking. The instrument can carry out period measurement of either 1 or 10 cycles in a choice of 6 time units; an a.c.-d.c. input switch allows high sensitivity on a.c. inputs while response is maintained on d.c. at a lower sensitivity.

The 1Mc/s crystal is maintained at a constant temperature by an oven, the resultant frequency being divided down to 0·1c/s by binary stages.

**EE 68 769** *for further details*

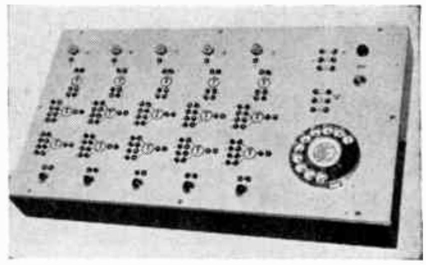
## EDUCATIONAL DIGITAL COMPUTER

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

*(Illustrated above right)*

The 'LAN-DEC' digital computer has been designed to provide training establishments, with an inexpensive high quality instrument capable of a wide range of computer circuit demonstrations.

Normally supplied for bench use the 'LAN-DEC' may also be fitted flush in a standard 19in rack simply by removing the side pieces.



The 'LAN-DEC' is basically a large patch board—19in × 10½in × 2½in with standard computer logic symbols representing each of the fifteen NOR gate transistor resistor logic circuits, which are wired directly to the underside in five vertical columns of three gates each. Connexions to each of the NOR gate inputs and outputs are made via high quality silver plated miniature sockets which are duplicated at each point.

Each computer is supplied complete with all necessary cords for patching etc, and a comprehensive manual covering a wide range of computer circuit demonstrations with patching instructions and drawings etc. Additional copies may be purchased separately.

Add-on relay control unit type RCU. 5 is available and provides control facilities for external mechanisms.

OC83 transistors are used throughout in the computer, and their ability to withstand up to 0·5A overload provides a much greater safety factor than is usual with instruments of this type.

'LAN-DEC' was designed with the co-operation of the Electrical Engineering and Mathematics Depts of Enfield College of Technology, and Edmonton County Grammar School.

**EE 68 770** *for further details*

## D.C. POWER SUPPLIES

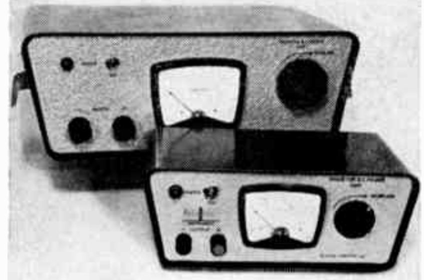
**De La Rue Frigistor Ltd, Canal Estate, Langley, Buckinghamshire**

*(Illustrated below)*

A range of low priced mains operated d.c. power supplies is announced by De La Rue Frigistor Ltd, a member of The De La Rue Group.

Each unit delivers a high current, low voltage d.c. supply, infinitely variable by manual control and smoothed so that ripple at any part of the range does not exceed 10 per cent.

Three standard cases, attractively styled, house eleven different units delivering 15 or 30A at 1½, 3, 6 and 9V or 60A up to 6V.



# TRIP AMPLIFIER

The Correx Tripamp is a magnetic amplifier driving a conventional relay at adjustable differential. It provides trip control on any varying monitor signal that is, or can be restored to, a direct current, and is adjustable to do so at any preset value within its operating range.

**VERSATILE**
Can be used over a wide range of applications, i.e. Thermocouples, strain gauges, photocells, load cells, resistance thermometer bulbs, contact thermometers, radiation pyrometers, etc.

**SENSITIVE**
Gives control at 0.025 mA minimum with a minimum input impedance of 60Ω.

**RELIABLE**
There are no moving or thermionic components apart from the relay which is type tested to $10^6$ operation at high frequency, plus the bonus feature of a rugged printed circuit. TRIPAMP is unaffected by vibration.

**COMPACT**
Designed to give maximum ease of servicing and fitting.

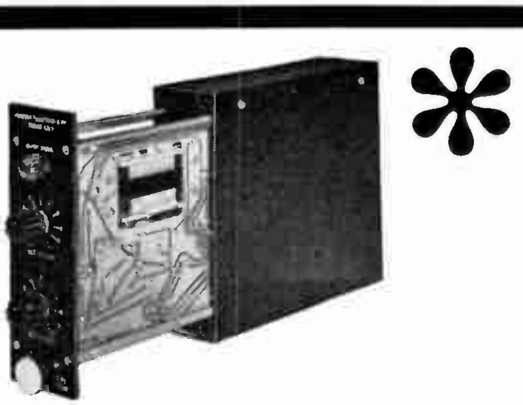
**ADJUSTMENT**
The differential is adjustable over wide limits.

**STABILITY**
Unaffected by variations in mains input within limits of ±8%.

**OUTPUT CONTACTS**
Single pole change-over is fitted as standard but provision is made for alternatives and up to four single pole change-overs can be fitted as optional extras. Alternatively TRIPAMP can be optionally supplied with Stabilised Voltage Reference for use with Slide-wire transducers.

**PRICE**
Basic model £19.15.0. List (As illustrated)

CORREX



## MODULAR EQUIPMENT

**FOR STEPLESS PROPORTIONAL CONTROL OF VOLTAGE, SPEED, POWER, HEAT, LIGHT, ETC.**

This range is readily available as individual units or as packaged systems designed to enable you to construct control schemes using the CORREX modular "add on" system.

**Included, for use with and in this range are:—**

**PROPORTIONAL CONTROLLERS**
**MAGNETIC INSTRUMENT AMPLIFIERS**
**MAGNETIC AMPLIFIER DRIVER STAGES**
**TRIP AMPLIFIERS (MAGNETIC)**
**SATURABLE REACTORS in a wide range UP TO 180 KVA**
**TRANSFORMERS**

**Other specialised control equipment for all purposes.**

**PHOENIX TELEPHONES LIMITED**
GROVE PARK, LONDON, N.W.9
Telephone COLindale 7243 Telegrams PHONOFENIX LONDON NW9

The standard unit includes control knob, ammeter, mains light and fuse. The units can be supplied at a considerably reduced price for building into equipment; these are supplied without control components, and deliver a predetermined d.c. output.

**EE 68 771** *for further details*

## D.F. RECEIVER

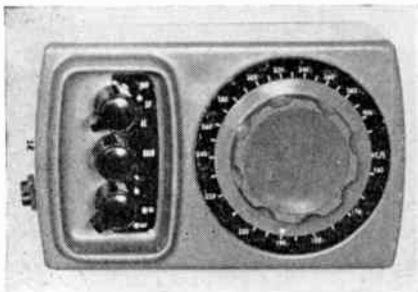
**Brookes & Gatehouse, Ltd, Bath Road, Lymington, Hampshire**

*(Illustrated below)*

Known as the 'Homer' this d.f. receiver which measures only 8¼ × 4⅞ × 3⅜in is intended for use in small yachts or fishing boats.

The Homer provides the following facilities:

(1) D.F. fixing when used with the Heron aerial. The frequency



coverage embraces all the maritime and aeronautical beacons, throughout the world. (190 to 410kc/s).

(2) Fixing by Consol and Consolan.

(3) With Model K: Reception of weather forecasts, shipping transmissions, etc., on the following bands:
  Long-wave (160 to 250kc/s).
  Medium-wave (550 to 1 600kc/s).
  Marine r.t. band (1 600 to 4 100kc/s).

With Model L: Reception of weather forecasts in European waters on the long-wave band (160 to 250kc/s). This model does not cover the medium and r.t. wavebands.

The Homer is contained in a sturdy, hermetically sealed case of aluminium alloy which is coated with nylon. The battery is contained in a separate compartment recessed into the back of the case which has its own sealing cover. It comprises four Mallory mercury cells, type ZM9, which are generally available in all countries (RM9 in U.S.A.), the endurance of which is about 250 hours of continuous operation. A bracket is supplied to enable the set to be mounted on a bulkhead or instrument panel. All fastenings are of stainless steel, control shafts are sealed by means of synthetic rubber glands and plugs and sockets are palladium-plated to prevent corrosion by sea water.

A wire aerial of height 15 to 30ft is used for broadcast and radiotelephony reception. The aerial input impedance is very low, enabling a wooden or plastic boat's standing rigging to be used as an aerial without the insertion of insulators. On long-wave, either the wire aerial or the Heron d.f. aerial may be used for broadcast and Consol/Consolan reception.

The Heron is a directional aerial which, in conjunction with a radio receiver enables fixes to be obtained from radio beacons by the method of intersecting lines of bearing. Using a principle which depends upon the use of a tuned ferrite rod, this instrument has the outstanding advantage of high accuracy and very small size.

In models A, B, C and D the Heron is integral with a compass, which permits bearings to be observed directly in relation to magnetic north. This arrangement avoids the errors due to wandering of the ship's head, and to heeling, which are inherent in conventional d.f. systems. For steel yachts, where compass deviation cannot be adequately corrected, a model E arrangement is available. This employs a conventional 360° azimuth scale and turntable and is gimballed to eliminate heeling error.

**EE 68 772** *for further details*

## GALVANOMETER DRIVE AMPLIFIER

**S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex**

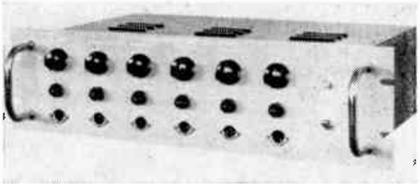
*(Illustrated above right)*

S. E. Laboratories (Engineering) Ltd have announced a transistorized drive amplifier type S.E. 425, designed to drive high-frequency galvanometers from low-power sources e.g. telemetering amplifiers, audio oscillators, playback discriminators, etc. Other applications include the measurement of small signals at high voltage-levels e.g. to check switching surges in electrical installations, motor starting circuits, etc.

Six electrically isolated amplifiers on plug-in printed-circuit cards are housed on a common chassis in a standard 19in rack-mounting cabinet.

Among the outstanding features of this amplifier are its very high signal-to-noise ratio of 1000 : 1; its low drift of less than ±0·5 per cent over a temperature-range of 0 to 45°C; and its wide frequency-response, which is flat within ±2 per cent between d.c. and 10kc/s. Interchannel crosstalk between the amplifiers is less than 2mV peak-to-peak compared with the full output of ±3·2V into a 50Ω load.

Input connexions to the amplifiers are electrically isolated from one another and can withstand up to 250V above ground. Inputs can be either balanced or unbalanced. Six individual three-pin plugs on the front panel are commoned with a multi-channel connector on the rear frame. When the amplifiers are used in conjunction with galvanometers they provide excellent common-mode rejection.



To protect galvanometers from excessive overloads, the amplifier output is limited at approximately 120mA. As with other S.E.L. electronic equipment, construction is rugged, and the circuits are well-tried and reliable.

**EE 68 773** *for further details*

## PORTABLE CHART RECORDER

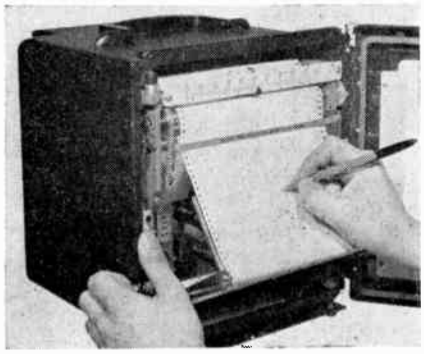
**Leeds & Northrup Ltd, Wharfdale Road, Tyseley, Birmingham, 11**

*(Illustrated below)*

Leeds & Northrup Ltd have developed a new version of the Speedomax 'H' range, Known as the Compact A.Z.A.R. (adjustable zero, adjustable range) this 6½in portable chart recorder has a wide application in research and testing laboratories, where a versatile instrument is required.

It employs independent zero and span controls individually powered by mercury cells; can provide any calibrated span between 0·3 and 50mV and any zero suppression between −50 and +50mV. With these features it is possible to plot a variety of measurements, such as temperature, stress, strain, position vibration and other e.m.f. variables, including gas chromatography.

A 'run-calibrate' switch is supplied to permit the setting of span and zero without the need of external calibrating equipment. Measuring circuits are designed to minimize current drains from the mercury cells to ensure long life expectancy. A span step-response of one second nominal is provided. It is possible to 'step' the span at less than 0·5μV and the zero at better than 0·0033 per cent of reading. Voltage stability with a new mercury cell is ±0·025 per cent of reading. Recommended options for the "Compact 'H' AZAR" include: fast and slow chart speed changer, note-taking tilting chart table and tear off bar, moderate high impedance amplifier for rated performance with external circuit impedance from 0 up to 18kΩ, feet and carrying handle.



**EE 68 774** *for further details*

# MEETINGS THIS MONTH

## INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1, unless otherwise stated.

**Radar Group**

Date: 1 April. Time: 6 p.m.
Short Paper: Pulse Compression Techniques in Radar.

**Computer Group**

Date: 9 April. Time: 6 p.m.
Lecture: Road and Rail Traffic Control using Computers.
By: H. A. Codd, A. Gilmour, D. G. W. Mace and E. J. Preston.

**Medical Electronics Group**

Date: 29 April. Time: 6 p.m.
Discussion: A Practical Approach to Transistor Circuits:

**Symposium**

Date: 15 April. Time: 10.30 a.m.
Held at: Birkbeck College, Malet Street, W.C.1.
Symposium: The Operation of Electronic Equipment under Conditions of Severe Electrical Interference.
Lecture: Surge Interference—its Importance and Cure.
By: D. S. Hiorns.
Lecture: Experiences in the Operation of Electronic Equipment on Heavy Industry Supply Systems.
By: G. B. Miller.
Lecture: The Frequency of Occurrence and the Magnitude of Short Duration Transients in Low Voltage Supply Systems.
By: J. H. Bull and W. Nethercot.
Lecture: Assessing Transient Voltages on Instrument Feeders.
By: H. J. Turner.
Lecture: The Mechanism of Interference Pick-up in Electronic Equipment with Special Reference to Nuclear Power Stations.
By: D. Harrison.
Lecture: Techniques for Reducing Interference Effects during Switchgear Tests in H.V. Sub-stations.
By: K. F. Foreman and J. N. Prewett.
(Advance registration will be necessary and application forms may be obtained from the Institution at 9 Bedford Square, W.C.1.)

**West Midlands Section**

Date: 30 April. Time: 7 p.m.
Held at: Electronic and Electrical Engineering Dept., University of Birmingham.
Lecture: Air Traffic Control.
By: D. H. Davies.
Date: 7 April.
Held at: University of Birmingham.
Symposium: Electronics in the Automobile Industry.
Further details and registration forms from Mr. G. K. Steel, College of Advanced Technology, Electrical Engineering Department, Gosta Green, Birmingham 4.

**South Western Section**

Date: 13 April. Time: 6.30 p.m.
Held at: Bristol University, Engineering Lecture Rooms, Bristol.
Lecture: Some Recent Ionosphere Researches.
By: J. A. Ratcliffe.
Date: 23 April. Time: 6.30 p.m.
Held at: Bristol College of Science and Technology.
Lecture: Microwave Spectroscopy and its Applications.
By: D. J. E. Ingram.

**North Eastern Section**

Date: 7 April. Time: 6 p.m.
Held at: Caterick Camp, Yorkshire.
Lecture: U.H.F. Propagation and Reception.
By: A. C. Roff.
Date: 8 April. Time: 6.15 p.m.
Held at: The Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle upon Tyne.
Annual General Meeting.
Lecture: Electronic Melody Instruments.
By: K. A. Macfadyen.

**Scottish Section**

Date: 3 to 5 April.
Held at: Heriot-Watt College, Edinburgh.
Symposium: Microminiaturization.
Registration will be necessary and application forms together with a copy of the programme may be obtained from the Honorary Secretary of the Brit. I.R.E. Scottish Section: Mr. R. D. Pittilo, 35 Crawfurd Road, Burnside, Rutherglen, Glasgow.

**South Midland Section**

Date: 3 April. Time: 7 p.m.
Held at: The B.B.C. Club, High Street, Evesham.
Lecture: Systems and Receiver Techniques for Stereophonic Broadcasting.
By: G. J. Phillips.
Held at: The Winter Gardens, Malvern.
Lecture: Superconductors in Instrumentation.
By: D. H. Parkinson.

**Merseyside Section**

Date: 15 April. Time: 7.30 p.m.
Held at: The Walker Art Gallery, Liverpool.
Lecture: The History of Radio.
By: G. R. M. Garratt.

**Southern Section.**

Date: 8 April. Time: 6.30 p.m.
Held at: Highbury Technical College, Cosham, Portsmouth.
Annual General Meeting.
Lecture: A Naval View of Reliability.
By: J. R. Young.

## INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m., unless otherwise stated.

**Ordinary Meeting**

Date: 30 April.
Lecture: Superconducting Magnets.
By: A. B. Pippard.

**Informal Meeting**

Date: 27 April.
Discussion: Developing the Telephone Service.
Opened by: A. Mumford.

**Electronics Division**

Date: 9 April.
Discussion: Road and Rail Traffic Control using Computers.
Opened by: H. A. Codd, A. Gilmore, D. G. W. Mace and E. J. Preston.
Date: 13 April.
Colloquium: Field-effect and other High-input-Impedance Devices and their Applications.
(Advance registration will be necessary and application should be made to the I.E.E. at Savoy Place, W.C.2.)
Date: 15 April.
Lecture: A New Method of Application using Acoustic Waves:
By: E. A. Ash.
Date: 16 April.
Colloquium: The Design and use of Cathode-Ray Tubes.
Date: 20 April.
Lecture: The Amplification of Oblique Incidence Sounding to Long-distance Communications Problems.
By: P. A. C. Morris.
Date: 21 April.
Colloquium: Simulators for Training Purposes.
(Advance registration will be necessary and application should be made to the I.E.E. at Savoy Place. W.C.2.)
Date: 29 April.
Colloquium: Filter design.
Date: 29 April.
Discussion: A Practical Approach to Transistor Circuits.

**Power Division**

Date: 1 April.
Discussion: Safety Considerations in System Design:
Opened by: L. A. Corney and J. Baskind.
Date: 14 April.
Lecture: Study of the Growth of Discharges on Polluted Insulation.
By: L. L. Alston and S. Zoledziowski.
Lecture: Artificial Pollution Test for High-Voltage Outdoor Insulation.
By: C. H. A. Ely and P. J. Lambeth.
Lecture: Flashover Mechanism of Polluted Insulation.
By: B. F. Hampton.
Date: 20 April.
Lecture: Modern Developments in Silicon-Iron (joint meeting with Science and General Division).
By: J. E. Thompson.
Date: 22 April.
Lecture: Railway Electrification in India.
By: H. D. Awasty.
Date: 29 April.
Lecture: Review of Marine a.c. Installations.
By: D. Gray.

**Science and General Division**

Date: 7 April.
Lecture: Electronic Summation Metering.
By: M. L. Done.
Date: 8-10 April.
Conference: Dielectric and Insulating Materials.
(All wishing to attend must register; forms available on application.)
Date: 9 April.
Discussion: Road and Rail Traffic Control using Computers (joint meeting with the Brit. I.R.E. Computer Group).
Opened by: H. A. Codd, A. Gilmore, D. G. W. Mace and E. J. Preston.
Date: 14 April.
Lecture: Power Generation using Radioactive Isotopes.
By: F. W. Yates.
Date: 16 April.
Discussion: The Fielden Report and the Eelectrical Engineer.
Opened by: R. L. Cannell.
Date: 21 April.
Discussion: Adaptive Control in Aircraft.
Opened by: W. H. McKinlay.
Date: 21 April. Time: 2.30 p.m.
Colloquium: Simulators for Training Purposes.
Date: 27 April.
Discussion: The Choice between Analogue and Digital-Computing Techniques in Engineering Analysis and Design.
Opened by: R. A. Laws, P. K. M'Pherson, R. J. A. Paul, and K. L. Smith.

## THE SOCIETY OF INSTRUMENT TECHNOLOGY

All meetings will be held at Manson House, 26 Portland Place, London, W.1.
Date: 8 April. Time: 5 p.m.
Symposium: Accurate Dispensing of Liquids below 100 ml.
Date: 28 April. Time: 7 p.m.
Lecture: The Use of Matrix Methods Applied to the Design of Multivariable Control Systems.
By: L. Finkelstein and J. G. Hargrave.

## THE TELEVISION SOCIETY

Date: 13 April. Time: 7 p.m.
Held at: The Faraday Room, I.E.E., Savoy Place, W.C.2.
Lecture: Problems of Recording Colour Signals on Magnetic Tape.
By: P. Rainger.

## THE INSTITUTION OF ENGINEERING DESIGNERS

Date: 13 April. Time: 7.15 p.m.
Held at: The New Lecture Theatre, Rutherford College of Technology, Northumberland Road, Newcastle upon Tyne 1.
Lecture: Air Filters and their Application for Air Cleaning and Sterilization.
By: J. E. Firman.

# PUBLICATIONS RECEIVED

EMI's NEW CAMERA TUBES booklet is now available. Step by step, the manufacture of a television camera tube is shown in this publication which is now available from EMI Electronics Ltd. From the mounting of the mesh to the final testing stage, the reader is able to look over the operatives' shoulders and see the 'eye' of a modern television camera gradually taking shape. Described in detail are the EMI's 4½in image orthicon and high resolution vidicon tubes. Copies of this booklet can be obtained from EMI Electronics Ltd, Valve Division, Hayes, Middlesex.

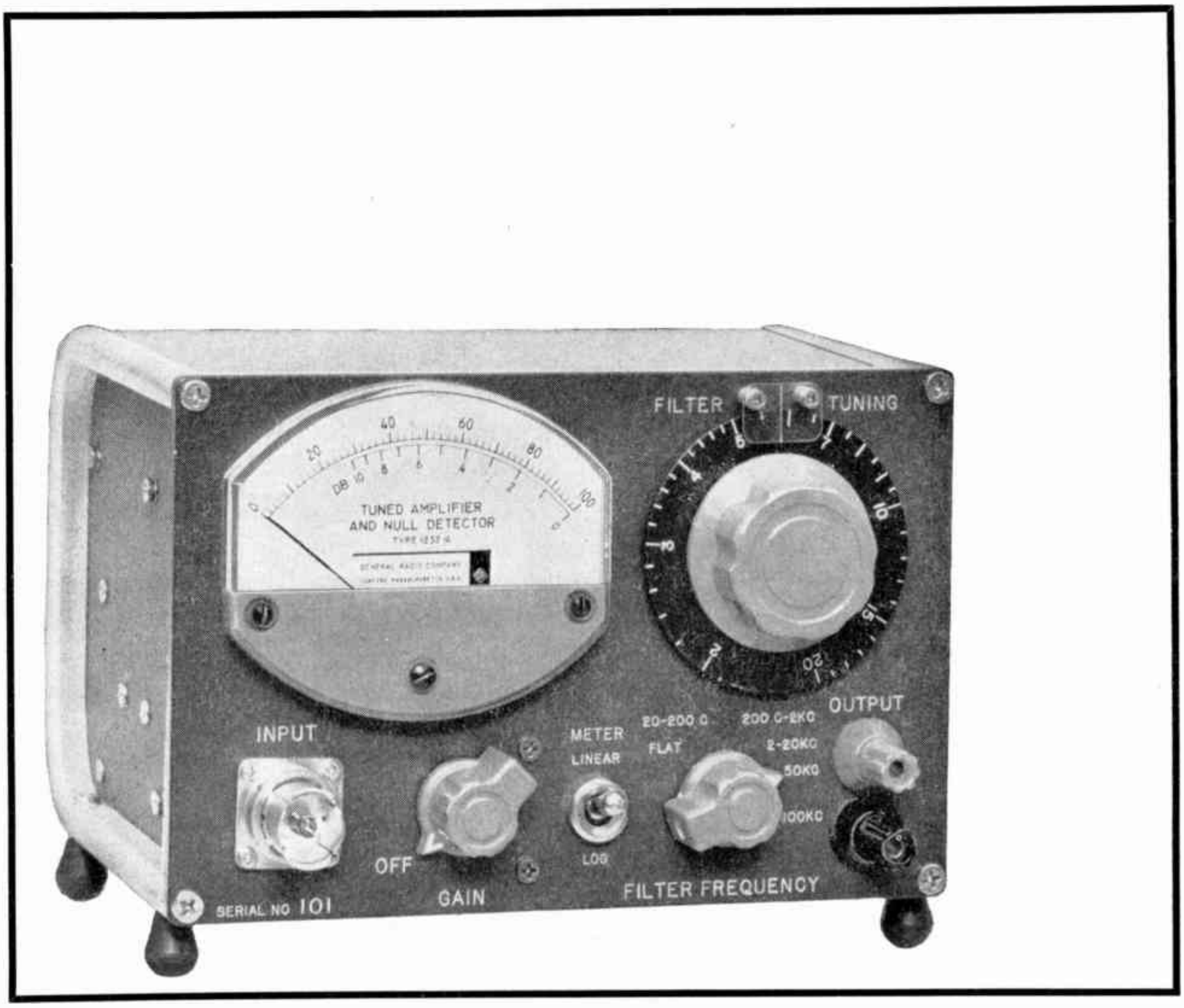
MULLARD COMPONENTS is the title of a quick reference guide which contains abridged information on all components available from Mullard, including two recently introduced ranges—Piezoelectric Materials and Recording Heads. Data Sheets giving full information on these products are published in the Mullard Technical Handbook or can be obtained individually on request. Requests for copies of the quick reference guide should be addressed to the Company, on headed company notepaper, at Mullard House, Components Division, Torrington Place, London, W.C.1.

PRECISION PRODUCTS is the title of a new publication by Muirhead & Co. Ltd, of Beckenham Kent. It contains details in English, French and German of precision electrical instruments, synchros and servo systems and Facsimile communication systems. Requests for copies should be addressed to the Company at the above address.

**EE 68 148** *for further details*

# AUDIO FREQUENCY TUNED AMPLIFIER AND NULL DETECTOR WITH **1μV FULL-SCALE SENSITIVITY**

**CLAUDE LYONS LIMITED**




*General Radio Type 1232-A Tuned Amplifier and Null Detector*

The type 1232-A is a transistorised, battery operated, audio frequency null detector, featuring unusually high sensitivity—typically 100 nV (0·1 μV) or better—low noise level, excellent selectivity and high gain. These characteristics make it suitable for the most exacting bridge measurement requirements, and also make it invaluable as an oscilloscope or transducer pre-amplifier or a sensitive wave analyser.

*As the exclusive U.K. representatives, Claude Lyons Ltd will be glad to supply full details of General Radio precision electronic measuring instruments.*

- High sensitivity — 1 μV *full scale*
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- Wide range — tunable 20 c/s – 20 kc/s plus fixed 50 kc/s and 100 kc/s positions
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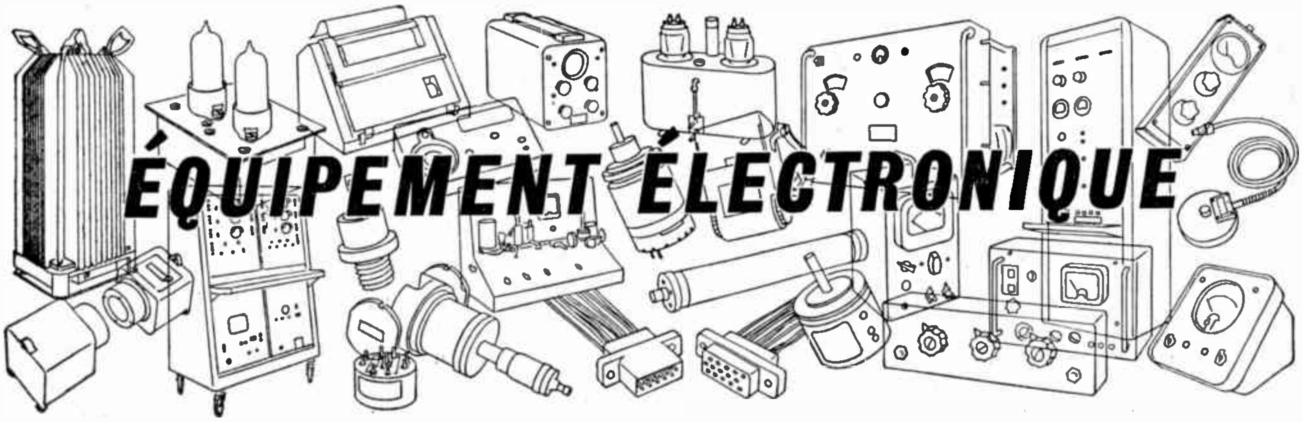


# ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

*Traduction des pages 268 à 273*

### ÉQUIPEMENT DE SONDAGE HALL

**Scientifica, 148 St. Dunston's Avenoe, Acton, London. W.3**

(*Illustration à la page 268*)

En utilisant l'effet Hall, cet équipement assure un moyen simple et pratique de mesurer la puissance de champ magnétique. On peut ainsi, en reliant directement la sortie de la sonde Hall à un enregistreur de 50 $\mu$A, mesurer des puissances de champ de 100 à 20 000 gauss, sans la moindre difficulté. L'extrémité inférieure de cette gamme peut, cependant, être étendue d'avantage par l'emploi d'un amplificateur à transistors à un seul étage dont la configuration du circuit est indiquée en détail dans le manuel d'instructions.

La sonde Hall est, en effet, fournie avec un manuel illustré contenant des graphiques faciles à suivre et qui indique la procédure élémentaire à suivre pour l'assemblage de la sonde. Il comporte également une description claire et détaillée des principes théoriques de l'appareil. Ce manuel permet donc aux étudiants d'exécuter de nombreuses expériences en utilisant l'effet Hall et, en même temps, d'appécier la signification pratique de cette découverte.

L'étalonnage de la sonde Hall s'effectue à l'aide d'un aimant normal fourni avec l'appareil. En variante, on peut employer l'aimant normal en liaison avec la sonde afin de donner des démonstrations de l'effet Hall ou encore pour mesurer les paramètres Hall.

La tension de sortie pour un champ de 10 kG est de 200 mV et la stabilité thermique est de 0,3%/°C.

**EE 68 751** *pour plus amples renseignements*

### INTERRUPTEUR DE TEMPORISATION À RUPTURE BRUSQUE

**Standard Telephones & Cables Ltd., Connaught Hoose, 63 Aldwych, London, W.C.2**

(*Illustration à la page 268*)

Un nouvel interrupteur de temporisation thermique à action rapide, modèle S45 C/ID, peut être obtenu maintenant de la Standard Telephones and Cables Ltd. Conçu pour la commutation de courants relativement élevés (jusqu'à 250 V à 5 A, c.a.), le nouvel interrupteur est un dispositif de commutation unipolaire à pression de contact élevée et à action 'repos-travail' rapide.

La durée nominale de commutation retardée est de 42 sec. En reliant un certain nombre de dispositifs en série, on peut obtenir des multiples de ce temps de retard.

Les interrupteurs de temporisation thermique sont particulièrement indiqués pour assurer le retard entre l'application de la tension de chauffage et les autres tensions de circuit, telle que la tension anodique, à des tubes et lampes indirectement chauffés. Ils peuvent, en outre être utilisés comme oscillateurs de relaxation à très basse fréquence, ainsi que pour la commutation de circuits triphasés de démarrage étoile-triangle pour moteurs à induction et pour refermer un rupteur de circuit après son déclenchement provoqué par une surtension temporaire.

**EE 68 752** *pour plus amples renseignements*

### MINUTERIE D'INTERVALLES DE PRÉCISION

**Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex**

(*Illustration à la page 268*)

Cette minuterie d'intervalles a été spécialement conçue pour mesurer en millisecondes les durées de fonctionnement de contacts de relais, mais elle peut être utilisée tout aussi bien pour déterminer n'importe quel intervalle de temps délimité par un ou plusieurs contacts extérieurs.

Des impulsions intérieures au rythme de 1000 par seconde, obtenues d'un oscillateur piloté par quartz (d'une précision de 0,01 %) sont comptées par quatre tubes Dékatron en cascade pendant la période de fonctionnement du contact ou des contacts extérieurs, le temps qui en résulte (jusqu'à un maximum de 9 999 sec) étant lu directement sur ces tubes Dékatron.

Le selecteur à six directions et les deux paires de bornes d'entrée à ressorts permettent d'arrêter ou de mettre en marche la minuterie par n'importe quelle combinaison d'un ou de deux contacts d'ouverture ou de fermeture. L'appareil fonctionne sur courant secteur alternatif de 200 à 250 V et il est muni d'un interrupteur de "réenclenchement-fonctionnement" à trois directions.

Il existe plusieurs modèles de rechange à comptage maximum de 999 99 sec (en centièmes de seconde) et tous les modèles peuvent être fournis avec ou sans mallette de transport en bois verni.

L'appareil peut aussi être prévu pour le fonctionnement par impulsions extérieures ou par tensions continues, moyennant un léger supplément de prix.

**EE 68 753** *pour plus amples renseignements*

**FOIRE INTERNATIONALE DE HANOVRE**

**"Electronic Engineering" occupera le Stand 2, dans la Salle 13, à la Foire Internationale de Hanovre, du 26 avril au 5 mai 1964, et les visiteurs y seront les bienvenus.**

### PONT À ÉQUILIBRAGE AUTOMATIQUE

**The Wayne Kerr Laboratories Ltd, New Malden, Sorrey**

(*Illustration à la page 268*)

Ce nouvel instrument, réalisé par The Wayne Kerr Laboratories Ltd, permet d'effectuer la mesure précise des facteurs résistifs et réactifs d'une impédance inconnue sans équilibrage manuel. Utilisé en liaison avec le pont universal Wayne Kerr type B221, le nouvel adaptateur à auto-équilibrage, type AA221, maintient électroniquement un état d'équilibre constant, indiquant immédiatement sur un cadran d'appareil à double enregistreur les variations dans la valeur des facteurs résistifs ou réactifs.

Les deux instruments de mesure de l'adaptateur à auto-équilibrage remplacent efficacement les commandes vernier sur le pont. Ainsi, par un choix approprié de la gamme du pont, les instruments de mesure peuvent fournir soit la lecture complète soit, avec les premiers deux chiffres fixés sur les commutateurs à décade du pont, une mesure globale à quatre chiffres significatifs près. Dans l'un ou l'autre de ces cas, les commandes vernier sont réglées à zéro et l'équilibre final est maintenu électroniquement. La précision de base de 0,1 %, la gamme exceptionnelle et la discrimination du pont B221 ne sont nullement affectés lorsqu'il est utilisé avec l'adaptateur à auto-équilibrage.

Pour le contrôle de lots, les commandes vernier du pont peuvent être réglées de manière à ce que les instruments de mesure puissent lire à mi-échelle à la valeur nominale voulue. En utilisant un simple masque sur les échelles des instruments de mesure, on peut contrôler des composants en employant une main d'oeuvre non spécialisée. Les sorties des circuits de mesure sont également reliées aux prises de jack à l'arrière de l'adaptateur d'auto-équilibrage pour l'utilisation du matériel de comptage de lots automatique. On peut aussi utiliser des voltmètres numériques ainsi que des mécanisme imprimeurs, des enregistreurs et des instruments de mesure à lecture à distance.

L'adaptateur à auto-équilibrage constitue tant une source de signaux qu'un détecteur transistorisé et alimenté par une batterie intérieure. La combinaison du pont et de l'adaptateur peut donc être utilisée lorsque le courant alternatif ne peut être obtenu. Pour l'emploi continu, un bloc d'alimentation secteur peut être fourni.

**EE 68 754** *pour plus amples renseignements*

## TRACEUR DE RÉPONSE DE FRÉQUENCE ACOUSTIQUE

**Distributeurs: Claude Lyons Ltd, Hoddesdon, Hertfordshire**

*(Illustration à la page 269)*

Réalisé par Costruzione Elettriche di D. Borsini, Ancône (Italie), le traceur de réponse de fréquence automatique TAR/61 est un analyseur de spectres à fréquence acoustique qui simplifie le contrôle des appareils suivants, ainsi que les traveux expérimentaux effectués sur eux: amplificateurs, filtres, enregistreurs, microphones, pickups, haut-parleurs et transducteurs, ainsi que les éléments de circuit en général.

L'instrument comprend essentiellement un oscillateur à battements à faible distorsion, couvrant la gamme de 0 à 20 kHz, qui est balayé de manière répétée par un élément de commande. La sortie à fréquence variable et à niveau constant de l'oscillateur à battements est appliquée, au moyen d'un amplificateur de puissance à atténuateurs étalonnés, au dispositif soumis à l'essai, et la sortie de ce dernier est renvoyée au traceur où elle est appliquée à un amplificateur logarithmique qui alimente le système de déviation vertical d'un tube cathodique à longue persistance. Le balayage horizontal du tube cathodique s'obtient par un potentiomètre très robuste à 180 plots, monté sur l'arbre d'entraînement de balayage de l'oscillateur à battements, produisant ainsi un graphique direct d'amplitude/fréquence.

Le micromètre oscilloscopique est étalonné verticalement en décibels et en millivolts, et horizontalement en fréquence. La persistance de la trace est suffisamment longue pour permettre de superposer des balayages successifs aux fins de comparaison directe, rendant ainsi immédiatement évident le résultat de tout réglage. Des courbes de réponse de référence ou limites d'étalonnage peuvent être marquées sur le micromètre, simplifiant ainsi considérablement le contrôle de la production.

Lorsqu'il y a lieu d'examiner des dispositifs à deux bornes, tels que microphones ou haut-parleurs, il faut, bien entendu, employer un transducteur correspondant pour compléter le système: un haut parleur pour produire l'excitation nécessaire au microphone, par exemple. Le transducteur secondaire peut produire des erreurs dans les résultats, et on a donc prévu le branchement d'un amplificateur de compression au circuit oscillateur, alimenté par un transducteur auxiliaire, afin de corriger toute non-linéarité.

**EE 68 755** *pour plus amples renseignements*

## ENREGISTREUR DE CONTRAINTE ET DE TEMPÉRATURE

**Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex**

*(Illustration à la page 269)*

Cet équipement portatif de concentration de données, se composant de deux éléments, a été réalisé sous contrat du Department of Scientific and Industrial Research et permet de mesurer la contrainte et la température de structures en béton précontraint.

Il peut mesurer jusqu'à cinquante canaux de chaque paramètre, la sortie étant présentée numériquement tant sous forme d'enregistrement imprimé que sur bande perforée.

La contrainte est mesurée par l'extensomètre à fil vibrant. Cette jauge est prévue pour être logée dans la structure en béton. La période de 100 cycles de vibration de fil est enregistrée et le rapport simple de loi quadratique entre la contrainte et la période ainsi mesuré facilite le calcul direct de la contrainte intérieure dans la structure.

La mesure de température s'obtient par des éléments de thermomètres à résistance, les résultats étant affichés sous forme de lectures directes dans l'échelle absolue de température.

L'équipement peut être utilisé sans surveillance et son programme de travail peut être étudié pour le fonctionnement automatique à des intervalles prédéterminés pouvant atteindre une durée de deux heures.

**EE 68 756** *pour plus amples renseignements*

## RÉSISTANCES À FILM MÉTALLIQUE

**Dubilier Condenser Co. (1925) Ltd, Ducon Works, Victoria Road, North Acton, London, W.3**

*(Illustration à la page 269)*

Les résistances à film métallique type ME sont livrables pour cinq puissances nominales utiles et pour des valeurs de résistance de 30 Ω à 10 MΩ sujettes à des limitations de gamme par ordre de puissance utile. La tolérance normale est de ±1 % mais des tolérances plus réduites de 0,5 %, 0,25 % et 0,1 % peuvent être obtenues en fonction du coefficient de température.

L'enveloppe extérieure des modèles à puissance utile réduite est un moulage spécial d'isolants plastiques à température élevée et résistant à l'humidité. Pour les puissances utiles de 1 et de 2 W, un tube de porcelaine à scellements d'extrémité hermétiques est employé.

Les cinq modèles ont un niveau de bruit très bas, indépendant de la valeur de résistance, normalement au-dessous de 0,1 μV/V.

Le coefficient de tension, lorsqu'il est mesuré entre $1/10^e$ de tension et la tension nominale totale, est inférieur à $5{,}10^{-6}$/V pour toutes les valeurs. Sur les résistances à valeur moyenne et à valeur réduite, le coefficient de tension est essentiellement zéro. Les données indiquent que des changements pendant une période d'entreposage d'un an à une température de salle normale ne peuvent être mesurées sur un équipement de laboratoire normal. Les résistances peuvent être fournies à cinq coefficients de température différents allant de $\pm 150{,}10^{-6}$ jusqu'au coefficient réduit de $\pm 15{,}10^{-6}/°$ C.

Les résistances ne sont pas inductives jusqu'à environ 100 MHz, sauf dans le cas des types à wattage élevé et à faible valeur. Elles constituent des composants de remplacement très utiles pour les types bobinés de précision dont la miniaturisation est de grande importance.

**EE 68 757** *pour plus amples renseignements*

## OSCILLATEUR À FRÉQUENCE DE BATTEMENTS

**Distributeurs: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1**

*(Illustration à la page 269)*

La société Livingston Laboratories Ltd fournit un oscillateur couvrant en une seule gamme toutes les bandes de fréquence acoustique et sub-acoustique. Il s'agit de l'oscillateur de fréquence de battements type HO32 construit par Radiometer du Danemark.

Cet instrument a une gamme de fréquence totale de zéro à 21 kHz, la gamme de zéro à 20,5 kHz étant couverte en un seul balayage. La puissance de sortie maxima est de 4W et elle est fournie par un transformateur de sortie ou par un atténuateur étalonné. Le transformateur comprend des prises pour l'équilibrage d'impédance de diverses charges avec choix d'une sortie de 15 Ω. L'atténuateur fournit des niveaux de sortie de 300 μV à 100 V. La distorsion harmonique est inférieure à 0,1 % pour des sorties d'atténuateur inférieures à 10 V. L'instrument se caractérise également par une bonne stabilité et réponse de fréquence et par son indicateur à oeil magique permettant de déterminer facilement le battement nul.

**EE 68 758** *pour plus amples renseignements*

## ENREGISTREUR D'ISOLEMENT

**Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex**

(*Illustration à la page 270*)

L'enregistreur d'isolement type 190 a été étudié pour la mesure directe de la résistance d'isolement et du courant de fuite. C'est un instrument portatif utilisant un inverseur à transistors pour produire les tensions d'essai à partir d'une batterie intérieure. Des tensions d'essai de 500 V, 250 V et 100 V sont prévues. L'emploi d'un instrument de mesure à échelle de 8,89 cm de longueur permet de mesurer une résistance d'isolement de 1 000 MΩ.

La commutation marche-repos par bouton-poussoir assure le fonctionnement instantané à l'aide d'une seule main et élimine toute perte inutile de la batterie. Un seul commutateur rotatif est utilisé pour le choix de la tension d'essai voulue et il comprend une position d'étalonnage pour fixer la tension de sortie à l'aide d'une commande préréglée. L'emploi d'une batterie au mercure constitue une source de puissance compacte et donne une durée de vie moyenne de plus de 2 500 opérations avec une longue durée d'entreposage et l'absence de corrosion.

La sortie est stabilisée électroniquement et le courant maximum est limité à 50 μA. Le circuit de limitation du courant réduit la tension d'essai proportionnellement à sa valeur totale dans des conditions de circuit ouvert à zéro sur l'échelle totale. Cette possibilité élimine virtuellement le risque de panne destructrice et permet d'utiliser l'instrument en toute sécurité pour mesurer le courant de fuite de redresseurs semiconducteurs, de diodes etc.

**EE 68 759** *pour plus amples renseignements*

## TRANSDUCTEURS DE PRESSION

**Intersonde Ltd, The Forum, High Street, Edgware, Middlesex**

(*Illustration à la page 270*)

La société Intersonde Ltd vient d'introduire les transducteurs de pression d'extensomètre à liaison type PR14 et PR15, couvrant des gammes de pression allant de zéro à 70 kg/cm² et de 0 à 700 kg/cm pour mesurer la pression liquide ou gazeuse, statique ou dynamique.

L'élément sensible à la pression, commun aux deux types, est un tube usiné avec précision dont les éléments de jauge de contrainte à deux fils sont bobinés et reliés dans un montage en pont à quatre bras. Les deux transducteurs ont une résistance de sortie de 350 Ω et produisent une sortie de 22 mV à une puissance nominale complète avec excitation de 15 V. La non-linéarité et l'erreur d'hystérésis combinées sont inférieures à 0,25 % de la gamme totale et les limites de température de régime s'étendent de −40° C à +120° C.

Le type PR14 est muni de bornes à broches et le type PR15 comporte un conducteur mobile intégré. Les deux transducteurs mesurent 125 mm de long sur 25,7 mm de diamètre.

**EE 68 760** *pour plus amples renseignements*

## MAGNÉTRON À IMPULSIONS ACCORDABLES

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

(*Illustration à la page 270*)

Le magnétron accordable 7008 a des caractéristiques électriques semblables à celles de la série 4J50A. Il a une puissance de sortie de pointe caractéristique de 220 kW et une gamme de fréquence de 8 500 à 9 600 MHz. Son cycle de régime est de 0,001 et sa largeur d'impulsion va de 0,2 à 2,75 μsec à un courant d'anode de pointe de 27,5 A.

Les particularités spéciales du 7008 sont les suivantes: puissance de sortie de pointe minima dans toute la gamme de fréquence s'élevant à 200 kW; caractéristique d'accord linéaire avec cadran à étalonnage précis et approximatif; le mécanisme d'accord compact a été spécialement prévu pour un jeu réduit; taux élevé de montée de tension; cathode spéciale de longue durée; décharge disruptive réduite ou nulle; construction de chauffage unique en son genre pour assurer une plus grande fiabilité; refroidissement par convection naturelle dans certaines conditions de fonctionnement et d'environnement.

**EE 68 761** *pour plus amples renseignements*

## 'BATTERIES PELTIER'

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

(*Illustration à la page 270*)

La société Mullard Ltd a réalisé trois batteries Peltier convenant pour une gamme étendue d'applications de refroidissement exigeant une 'source froide' compacte.

Pour les utilisation où l'espace disponible est limité, la batterie Peltier offre des avantages considérables par rapport aux appareils de refroidissement de type classique car elle n'exige ni compresseur ni élément de chauffage. Ses applications caractéristiques sont le refroidissement et la réfrigération réduite pour la recherche médicale et biologique et pour les sources froides de matériel transistorisé.

Ces batteries dont l'élément thermoélectrique est constitué par du tellurure de bismuth fonctionnent suivant le principe Peltier, c'est à dire qu'un courant continu, passant à travers une jonction de métaux dissemblables ou de matériaux semiconducteurs, établit un gradient de température à travers cette jonction.

La batterie type PT11/20 a été prévue pour un courant de fonctionnement de 18 à 22 A à une tension de 1,0 à 1,2 V et sa capacité de refroidissement maximum est de 16 W. Sa durée de vie minima est de 2 000 heures de fonctionnement continu à 20 A.

Le type PT20/20 a un courant de fonctionnement de 20 A à 2 V, une capacité de refroidissement de 23 W et une durée de vie minima de 2 000 heures de fonctionnement continu à 20 A.

Le type PT47/5 fonctionne à un courant de 5 à 6 A et à une tension de 5,0 à 5,4 V et sa capacité de refroidissement est de 16 W. Sa durée de vie minima est de 2 000 heures de fonctionnement continu à 5 A. Tant le type PT11/20 que le type PT47/5 sont livrables avec plaque de cuivre pour surfaces solides, ou avec des ailettes pour le refroidissement des gaz ou des liquides. Le type PT20/20 a des plaques en cuivre.

Les trois modèles sont livrés prêts à l'emploi immédiat.

**EE 68 762** *pour plus amples renseignements*

## RÉFLECTOMÈTRES COAXIAUX

**A. T. & E. (Bridgnorth) Ltd, Bridgnorth, Shropshire**

(*Illustration à la page 271*)

La société A. T. & E. (Bridgnorth) Limited, du Groupe Plessey, a réalisé une nouvelle gamme de réflectomètres coaxiaux pouvant être installés de façon permanente dans les lignes coaxiales pour le contrôle de la puissance incidente réfléchie, pour les utilisateurs de matériel u.h.f.

Ces appareils se distinguent par leur capacité d'acheminement de courants élevés et leur affaiblissement d'insertion réduit. Grâce à l'emploi de coupleurs directionnels à cadre résistif dos-à-dos la mesure par lignes à fentes est rendue superflue.

Les modèles disponibles couvrent les gammes de fréquence de 50 à 500 MHz. Le modèle prévu pour la gamme de 150 à 300 MHz a une directivité minima de 40 dB. La sensibilité de couplage est de 500 μA pour une puissance incidente de 10 W. Le taux d'ondes stationnaires de ligne primaire maximum est supérieur à 1,05:1 et l'affaiblissement d'insertion de ligne primaire est négligeable. Le pouvoir d'acheminement de puissance est de 300 W.

**EE 68 763** *pour plus amples renseignements*

## POTENTIOMÈTRES DE PRÉCISION

**Salford Electrical Instruments Ltd, Peel Works, Salford, 3, Lancashire**

*(Illustration à la page 271)*

La société Salford Electrical Instruments Ltd a étendu sa gamme de potentiomètres de précision qui comprennent maintenant des modèles du format 11. Toutes les dimensions comprennent des ergots d'emplacement conformes aux conditions de cette grandeur internationale.

Des valeurs de résistance de 500 Ω à 50 kΩ peuvent être assurées, les potentiomètres étant bobinés sur mandrins toroïdaux à usinage de précision utilisant un fil de résistance nickel-chrome ou un fil en métal précieux pour les applications de faible bruit.

La gamme de température s'étend de −40° C à +85° C, la tension nominale étant de 2 W à 20° C. La tolérance de linéarité indépendante est de ±0,5 %.

Le potentiomètre est fourni sous forme jumelée, comprenant jusqu'à quatre pièces jumelées. Des prises supplémentaires peuvent être fixées, chacune d'elles étant reliée à un seul tour de fil. Les bornes latérales dorées constituent des pièces standard du potentiomètre, destinées à faciliter le soudage des connexions extérieures.

Des contacts doubles à souder en métal précieux sont utilisés tant sur la piste de résistance que sur le collecteur rhodié.

Le pivot en acier inoxydable est monté dans une cage à billes blindée de haute qualité. Un alliage d'aluminium antirouille de qualité supérieure est usiné de manière former un boîtier de 25,55 mm de diamètre.

**EE 68 764** *pour plus amples renseignements*

## TROUSSE À OUTILS 'PRESTINCERT'

**Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex**

*(Illustration à la page 271)*

La société Belling & Lee Ltd vend maintenant une nouvelle trousse à outils pour applications électroniques, électriques et de mécanique légère.

La gamme de composants Prestincert comprend des douilles filetées à autopoinçonnage et des bornes pour montage en une seule opération sur tôles ainsi que sur le nombreuses matières synthétiques d'isolement. Nul forage ou rivetage est nécessaire, et la pièce est pratiquement indestructible en cours d'usage normal.

La nouvelle trousse a été conçue pour répondre à la demande croissante des réalisateurs et techniciens qui veulent pouvoir utiliser les méthodes d'atelier durant les travaux de réalisation et faire usage des composants pratiques Prestincert dans leurs plans pour les chaînes de montage futures.

La trousse Prestincert comprend, en plus d'une petite presse poids léger développant jusqu'à 79 kg/cm², six conteneurs de composants Prestincert aux dimensions les plus couramment employées dans les industries mécaniques légères.

Les manchons filetés ont les trois pas réels suivants: 0,53 mm, 0,66 mm et 0,81 mm. Trois types de postes terminaux sont prévus pour les applications électriques et électroniques. Des quantités supplémentaires de tous les composants peuvent être fournies sur demande.

Six paires d'outils d'insertion, à codage de couleurs suivant les composants Prestincert appropriés, sont également compris dans la trousse. Ces outils peuvent être insérés dans la presse sur simple pression et ils sont retenus par une pince à ressort faisant corps avec eux.

Les outils, la presse et les composants Prestincert sont contenus dans une solide mallette de transport.

**EE 68 765** *pour plus amples renseignements*

## TÊTES D'ENREGISTREMENT MAGNÉTIQUES

**EMI Electronics Ltd, Hayes, Middlesex**

*(Illustration à la page 271)*

Une nouvelle série de têtes magnétiques de précision, à savoir la série Emidata 'E,' vient d'être mise au point par la société EMI Electronics Ltd. Ces têtes ont été spécifiquement conçues pour l'instrumentation multipistes et les systèmes de traitement de données exigeant une grande uniformité de rendement.

Les caractéristiques de fonctionnement remarquables et la longue durée de toutes les têtes d'un entassement sont obtenues grâce au contrôle rigoureux de la profondeur de l'entrefer au cours de la fabrication. Les têtes ont également un alignement d'entrefer et d'azimuth d'une précision plus élevée que celle de la gamme précédente de têtes de la série "S" avec lesquelles elles sont entièrement interchangeables.

L'azimuth moyen et la dispersion d'entrefer sont tels que les bords en ligne de tous les entrefers d'un assemblage se trouvent dans une zone de 5 μm de large et carrée par rapport à la base. L'élévation ou pente de l'empilage de tête est maintenu dans une zone semblable de 12.7 μm de large. Les longueurs d'entrefer standard sont de 12,7 μm pour l'enregistrement et de 5 μm pour le réenregistrement.

Les têtes donnent une réponse de fréquence élevée allant jusqu'à 100 kHz et leur réponse de longueur d'onde longue est de 2,5 mm. Le rejet diaphonique est extrêmement bon. Les têtes enregistreuses inscrivent des pistes plus larges que les têtes de réenregistrement, afin de tenir compte d'un déplacement éventuel de la bande.

Toutes les configurations de piste I.R.I.G. et S.B.A.C. sont prévues jusqu'à 33 pistes entrelacées sur bande de 2,54 cm. Un pas et une largeur spéciale de piste peuvent être prévus sur commande spéciale. Let têtes sont fournies bobinées à des fiches miniature Cannon.

**EE 68 766** *pour plus amples renseignements*

## SERVO-AMPLIFICATEUR MINIATURE

**Evershed & Vignoles Ltd, Acton Lane, Chiswick, London, W.4**

*(Illustration à la page 271)*

L'amplificateur à transistors miniature Evershed type PA.86T a été conçu pour l'entraînement de servomoteurs à phase partagée c.a. à des fréquences allant de 50 à 400 Hz, donnant une sortie maxima de 10 W à 20-0-20 V dans une impédance de 35 Ω par enroulement.

L'amplificateur est à cinq transistors, les composants étant montés à l'intérieur de la source froide de transistors de sortie. L'encapsulage complet dans le caoutchouc de silicium est utilisé afin d'exclure la crasse et l'humidité, et afin de permettre à l'amplificateur de résister à des chocs mécaniques considérables. Toutes les connexions sont reliées aux deux barrettes à vis sur le haut de l'amplificateur.

Pour le fonctionnement sur alimentation à courant alternatif, l'amplificateur doit être utilisé avec un transformateur d'entrée Evershed type PA.43597 à valeur nominale de sortie de 23-023 V et pouvant fournir le courant nécessaire à trois amplificateurs. Des enroulements supplémentaires sont prévus pour phases de référence de moteur et de générateurs, ainsi que pour transformateur interrupteur, le cas échéant.

**EE 68 767** *pour plus amples renseignements*

## OUTIL DE DESSOUDAGE

**Charles Austen Pumps Ltd, Petersham Works, High Road, Byfleet, Surrey**

*(Illustration à la page 272)*

Le dessoudage et l'élimination de composants défectueux dans les circuits imprimés peuvent présenter de sérieuses difficultés, surtout si ces défectuosités se rapportent aux broches multiples. Lorsqu'un grand fer à souder est utilisé, les composants adjacents sont affectés par la chaleur qui peut endommager l'élément de manière à le rendre irréparable de façon économique. Des court-circuits se produisent parfois en raison du déplacement d'un excès de soudure.

En raison de l'emploi accru des circuits imprimés, la société Charles Austen Pumps Ltd a réalisé le "Soldermaster" Austen, spécialement conçu pour surmonter ces difficultés. Il s'agit d'un petit fer à souder à température élevée comportant une mèche creuse à travers laquelle le vide peut être appliqué sur commande de l'opérateur au moyen d'une pompe. Le joint de soudure fondu par le fer est aspiré à travers la mèche et enfermé dans un réservoir. Les vapeurs de flux indésirables sont également extraites du voisinage du joint. La pompe est protégée par un filtre à élément poreux pouvant être nettoyé. La soudure solidifiée peut être retirée du réservoir à la fin de l'opération.

L'appareil se compose d'une pompe Austin "Capex" avec microfiltre Austen, d'un réservoir de soudure et tuyaux, d'un fer à souder de 50 W avec mèche creuse et d'une mèche de réserve. Il est fourni monté sur base, avec support de fer à souder et poignée de transport. Il mesure 18 cm × 15,24 cm × 15,42 cm et son poids total est de 2,721 kg.

**EE 68 768** *pour plus amples renseignements*

## ENREGISTREUR DE FRÉQUENCE DE TEMPS

**Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey**

*(Illustration à la page 272)*

La société Venner Electronics Ltd vient de créer le TSA3436, un instrument universel compact donnant une indication à six chiffres de fréquence et de période jusqu'à 1,2 MHz et assurant la mesure de temps de 1 µsec à 100 000 sec. Il peut également être utilisé comme compteur à des vitesses pouvant atteindre 1 Mp/s.

Conçu pour les travaux de laboratoire ou pour le contrôle de qualité dans l'industrie, cet appareil portatif entièrement transistorisé mesure 36,19 cm × 9,84 cm × 26,67 cm. Il se distingue particulièrement par trois temps de déclenchement, la suppression facultative entre les résultats et l'auto-vérification. Il peut effectuer la mesure de périodes de 1 ou de 10 cycles, avec un choix de six unités de temps. L'interrupteur d'entrée c.a.—c.c. permet une sensibilité élevée sur les entrées de tension alternative, cependant que la réponse est maintenue sur la tension continue à une sensibilité inférieure.

Le cristal de 1 MHz est maintenu à une température constante au moyen d'un four, la fréquence qui en résulte étant divisée jusqu'à 0,1 Hz par étages binaires.

**EE 68 769** *pour plus amples renseignements*

## CALCULATRICE NUMÉRIQUE D'INSTRUCTION

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

*(Illustration à la page 272)*

La calculatrice numérique LAN-DEC a été conçue afin de pourvoir les établissements de formation d'un instrument de haute qualité à bon marché et pouvant servir à une gamme étendue de démonstrations de circuits de calculatrice.

Normalement fourni pour banc d'essai, le LAN-DEC peut être également encastré dans un bâti standard de 48,26 cm en enlevant simplement les pièces latérales.

Le LAN-DEC consiste essentiellement en un grand panneau de préaffichage de 48,26 cm × 26,67 cm × 6,35 cm comportant des symboles logiques de calculatrice standard représentant chacun des quinze circuits logiques de résistance à transistors de porte NON, qui sont reliés directement à la partie inférieure en cinq colonnes verticales de trois portes chacune. Les connexions de chacune des entrées et sorties de la porte NON sont effectuées à l'aide de douilles miniature argentées de haute qualité, doublées à chaque point.

Chacune des calculatrices est fournie complète avec tous les fils nécessaires au préaffichage et un manuel complet indiquant une série étendue de démonstrations de circuits de calculatrice avec instructions de préaffichage, dessins, etc. Des exemplaires supplémentaires peuvent être achetés séparément.

L'élément de commande de relais type RCU.5, qui peut être monté accessoirement, permet de commander des mécanismes extérieurs.

La calculatrice est entièrement équipée de transistors OC83 et leur propriété de résistance à des surcharges maxima de 0,5 A assure un facteur de sécurité beaucoup plus élevé que celui de la plupart des instruments de ce type.

Le LAN-DEC a été réalisé en collaboration avec le Département de Mathématiques et d'Electricité du Enfield College of Technology et de la Edmonton County Grammar School.

**EE 68 770** *pour plus amples renseignements*

## BLOCS D'ALIMENTATION EN CONTINU

**De La Rue Frigistor Ltd, Canal Estate, Langley, Buckinghamshire**

*(Illustration à la page 272)*

La société De La Rue Frigistor Ltd, du Groupe De La Rue, a créé une gamme de blocs d'alimentation encontinu à bas prix.

Chacun de ces blocs fournit une alimentation en courant continu élevé à faible tension, pouvant être varié indéfiniment par commande manuelle et uniformisé de manière à ce que l'ondulation à n'importe quelle partie de la gamme ne dépasse pas 10 %.

Trois coffrets standard, d'une présentation séduisante, logent onze éléments différents fournissant 15 ou 30 A à 1½, 3, 6 et 9 V ou 60 A jusqu'à 6 V.

L'élément standard comprend le bouton de commande, l'ampèremètre, la lumière secteur et un fusible. Des éléments peuvent être fournis à un prix considérablement réduit lorsqu'ils sont destinés à être incorporés à une installation. Ces derniers sont livrés sans composants de commande et fournissent une sortie prédéterminée de courant continu.

**EE 68 771** *pour plus amples renseignements*

## RÉCEPTEUR RADIOGONIOMÈTRE

**Brookes & Gatehouse, Ltd, Bath Road, Lymington, Hampshire**

*(Illustration à la page 273)*

Ce récepteur radiogoniomètre, appelé "Homer," ne mesure que 21 cm × 12,38 cm × 8,57 et il a été conçu pour les petits yachts ou les bâteaux de pêche.

Le 'Homer' offre les avantages suivants:

(1) Montage radiogoniométrique avec l'antenne Heron. La fréquence couverte englobe toutes les balises maritimes et aéronautiques du monde entier.

(2) Montage par 'Consol' et 'Consolan.'

(3) Avec le modèle K: réception des bulletins météorologiques, des émissions de navigation, etc. Sur les bandes suivantes:
 Ondes longues (160 à 250 kHz)
 Ondes moyennes (550 à 1600 kHz)
 Bande radiotéléphonique maritime (1600 à 4100 kHz).
 Avec le modèle L: réception des bulletins météorologiques dans les eaux européennes sur la bande des ondes longues (160 à 250 kHz). Ce modèle ne couvre pas les ondes moyennes et radiotéléphoniques.

Le Homer est logé dans un robuste coffret en alliage d'aluminium, hermétiquement scellé et revêtu de nylon. La batterie est placée dans un compartiment à part à l'arrière du coffret et comporte son propre couvercle de scellement. Elle comprend quatre cellules au mercure Mallory, type ZM9, qui s'obtiennent généralement dans le monde entier (RM9 aux Etats-Unis). L'endurance de ces cellules est d'environ 250 heures de fonctionnement continu. Un support est fourni pour permettre de monter l'appareil sur une cloison ou un panneau à instruments. Toutes les attaches sont en acier inoxydable; les axes de commande sont scellées au moyen de presse-étoupe en caoutchouc synthétique, tandis que les fiches et douilles sont doublées au palladium afin d'empêcher la corrosion par l'eau de mer.

Une antenne de fil métallique d'une hauteur de 4,5 m à 9 m est utilisée pour la réception d'émissions de radiodiffusion et de radiotéléphonie. L'impédance d'entrée de l'antenne est très faible, permettant d'employer comme antenne le gréement en bois ou en matière plastique d'un petit bâteau, sans qu'il y ait lieu d'insérer des isolateurs. Sur ondes longues, on peut employer soit l'antenne en fil métallique soit l'antenne radiotéléphonique Heron pour la réception d'émissions de radiodiffusion et de Consol/Consolan.

L'antenne Heron est une antenne directionnelle qui permet d'obtenir, en liaison avec un récepteur radio, des indications émanant de radio-balises par la méthode d'intersection de lignes de relèvement. Grâce à l'application du principe basé sur l'emploi d'une tige de ferrite accordée, cet instrument acquiert l'avantage exceptionnel d'une grande précision et d'un encombrement fort réduit.

Dans les modèles A, B, C et D, l'antenne Heron fait corps avec le compas, ce qui permet de faire le point directement par rapport au nord magnétique. Ce dispositif évite les erreurs dues à la déviation du nez du navire, ainsi qu'à la bande, erreurs inhérentes aux systèmes de radiogoniométrie classiques. Pour les yachts en acier, dont la déviation du compas ne peut être corrigée dans une mesure suffisante, le dispositif 'E' est prévu. Ce dernier utilise une échelle azimutale de 360° et une table tournante et il est muni d'une suspension à la cardan afin d'éliminer les erreurs de gîte.

**EE 68 772** *pour plus amples renseignements*

### AMPLIFICATEUR D'ENTRAÎNEMENT DE GALVANOMÈTRE

**S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex**

(*Illustration à la page 273*)

La société S. E. Laboratories (Engineering) Ltd vient d'annoncer la réalisation d'un amplificateur de commande transistorisé, type S.E.425, conçu pour la commande de galvanomètres haute fréquence à partir de sources de faible puissance telles que les amplificateurs de télémesure, les oscillateurs acoustiques, les discriminateurs de réenregistrement, etc. Il peut être utilisé en outre pour la mesure de faibles signaux à des niveaux de tension élevés: vérification des surtensions de commutation dans les installations électriques, les circuits d'allumage de moteurs, etc.

Six amplificateurs électriquement isolés sur plaquettes de circuit imprimé à fiches sont logés sur un châssis commun dans un coffret standard de 48,26 cm pour bâti.

Les caractéristiques exceptionnelles de cet amplificateur comprennent entre autres, un rapport signal-bruit très élevé, soit 1 000 : 1, une dérive inférieure à ±0,5 % dans une gamme de température de 0 à 45° C, une réponse de fréquence étendue et linéaire à ±2 % entre le courant continu et 10 kHz. La diaphonie entre canaux d'amplificateurs est inférieure à 2 mV crête à crête par rapport à la sortie totale de ±3,2 V dans une charge de 50 Ω.

Les connexions d'entrée aux amplificateurs sont électriquement isolées l'une de l'autre et peuvent résister jusqu'à 250 V au-dessus du sol. Les entrées peuvent être équilibrées ou déséquilibrées. Six fiches individuelles à trois broches sur le panneau frontal sont reliées à un conducteur multivoies sur le cadre arrière. Lorsque les amplificateurs sont utilisés en liaison avec les galvanomètres, ils assurent un excellent rejet de mode commun.

Pour protéger les galvanomètres contre les surcharges excessives, la sortie de l'amplificateur est limitée à environ 120 mA. A l'instar de tous les autres appareils électroniques S.E.L., les amplificateurs sont de construction robuste et les circuits sont d'une réalisation sûre et éprouvée.

**EE 68 773** *pour plus amples renseignements*

### ENREGISTREUR À BANDE PORTATIF

**Leeds & Northrup Ltd, Wharfdale Road, Tyseley, Birmingham, 11**

(*Illustration à la page 273*)

La société Leeds & Northrup Ltd a mis au point une nouvelle version de la gamme Speedomax 'H.' Ce nouvel enregistreur à bande portatif de 16,51 cm, appelé Compact A.Z.A.R., (à zéro réglable et à gamme réglable) trouve de nombreuses applications dans les laboratoires d'essais et de recherche exigeant un instrument d'une grande souplesse d'emploi.

Le nouvel enregistreur comporte des commandes indépendantes de zéro et de portée, actionnées individuellement par des cellules au mercure. Il peut assurer n'importe quelle portée étalonnée entre 0,3 et 50 mV et n'importe quelle suppresion de zéro entre − 50 et + 50 mV. Grâce à ces caractéristiques, on peut effectuer une variété étendue d'enregistrements de mesure comme ceux de la température, de la contrainte, de l'effort, des vibrations de position et d'autres variables de force électromagnétique, y compris la chromatographie du gaz.

Un commutateur "marche-étalonnage" est prévu pour le réglage de portée et de zéro sans faire usage d'un appareil d'étalonnage extérieur. Les circuits de mesure sont conçus de façon à minimiser les pertes de courant des cellules de mercure et à garantir une longue durée de fonctionnement. Une réponse à plots de portée d'une seconde nominale est également prévue. On peut "graduer" la portée à moins de 0,5 μV et le zéro à plus de 0,0033 % de la lecture. La stabilité de la tension avec une nouvelle cellule de mercure est de ±0,025 de la lecture. Les options recommandées pour le "Compact 'H' AZAR" comprennent: un changeur de vitesse de papier rapide et lent, une table basculante pour la prise de notes et une barre d'arrachement, un amplificateur d'impédance élevée moyenne pour performance nominale, avec impédance à circuit extérieur de 0 à 18 kΩ, des pieds et une poignée de transport.

**EE 68 774** *pour plus amples renseignements*

# Résumés des Principaux Articles

**Lecteur de bande et analyseur statistique** par J. A. Phillips

*Résumé de l'article aux pages 216 à 220*

*L'auteur décrit un appareil capable de lire les rouleaux de bandes et d'en tirer la lecture moyenne, la variance et un histogramme. Le système d'exploration est à tube de télévision vidicon. Il est entièrement automatique et donne une sortie codée binaire. Une petite calculatrice à programme fixe est utilisée pour les calculs de statistique.*

**Un circuit transistorisé pour tube éclair Xénon** par P. G. M. Dawe

*Résumé de l'article aux pages 220 à 221*

*L'auteur décrit un circuit pouvant allumer un tube éclair Xénon au moyen d'un interrupteur à transistors, actionné par des impulsions obtenues d'un flip-flop et d'un multivibrateur. La commande à transistors se caractérise par l'application d'un circuit ionisant auxiliaire à la lampe afin d'y maintenir une voie de décharge continue.*

**Antennes d'exploration de fréquence** par M. F. Radford

*Résumé de l'article aux pages 222 à 226*

*L'antenne d'exploration de fréquence est une antenne conçue de manière à ce que la direction du faisceau rayonné soit fonction de la fréquence. Cet article expose les principes généraux de ces antennes et montre qu'elles peuvent constituer une variante économique des antennes mécaniquement actionnées.*

**Quelques circuits simples de couplage à Dékatron** par A. J. Oxley

Résumé de l'article aux pages 227 à 232

*Quatre circuits de couplage à courants porteurs série et en parallèle, dont trois se prêtent au comptage réversible, sont décrits dans cet article. Les vitesses de comptage maxima sont indiquées pour chacun de ces circuits; elles sont toutes inférieures à la limite imposée par le tube Dékatron. Il est montré, cependant, que des circuits de couplage simples peuvent être utilisés en certains cas.*

**Un nouveau type de multivibrateur à marche libre** par V. M. Ristic

Résumé de l'article aux pages 232 à 233

*Cet article analyse un multivibrateur accouplé à un émetteur du type série et utilisant deux transistors pnp. Le circuit présente des temps de montée et de chute excellents et il oscille entre 0,01Hz et 250kHz.*

**Une nouvelle méthode pour reconnaître la résistance en cas de réactance** par E. R. Wigan

Résumé de l'article aux pages 238 à 239

*Il s'agit ici d'une technique du type à pont qui permet de déterminer la résistance et, au besoin, de la mesurer en cas de réactance. La base de mesure théorique est indiquée en détail mais un montage précis de circuit n'est pas décrit.*

**Un amplificateur à transistors avec stabilisation de réaction du courant continu et du courant alternatif** par H. C. Bertoya

Résumé de l'article aux pages 240 à 245

*L'auteur décrit et analyse un amplificateur à transistors utilisant une boucle de réaction de courant continu pour stabiliser le point de régime de courant continu. Si la proportion voulue de réaction de courant alternatif ne coincide pas avec la proportion de réaction de courant continu, on peut avoir deux boucles de réaction séparées.*

*Le couplage direct est utilisé entre les étages, ce qui donne une réalisation simple et économique.*

*Des exemples pratiques sont donnés d'emploi de transistors pnp aussi bien que de transistors npn. Un de ces exemples a été spécifiquement prévu pour les applications à bruit réduit.*

**Méthode perfectionnée pour mesurer la résistance de la peau** par L. E. H. Lindahl, A. H. Christianson, et R. N. Gooch

Résumé de l'article aux pages 245 à 246

*Cet article traite d'une nouvelle méthode pour mesurer la résistance cutanée. L'amplificateur de courant continu équilibré utilisé obvie à la non-linéarité, au manque de sensibilité et à la sortie insuffisante de bon nombre des instruments actuels. En outre, il assure l'étalonnage automatique. L'expérience a démontré que sa stabilité est bien en deçà des limites de tolérance nécessaires aux travaux psycho-physiologiques.*

**Redresseurs à avalanche au silicium** par G. Duddridge

Résumé de l'article aux pages 247 à 249

*La fabrication et les caractéristiques des redresseurs à avalanche au silicium sont décrites dans cet article qui relève, en outre, le fait que ce type de redresseur a le pouvoir d'absorber des transitoires de surtension de réserve. L'article décrit le fonctionnement en parallèle et en série et indique quelques applications.*

**Un isolateur conique de résonance de champ à large bande** par J. H. Collins

Résumé de l'article aux pages 250 à 252

*Cet article décrit les travaux effectués en vue de réaliser un isolateur de résonance à large bande utilisant un champ magnétique de courant continu extérieur à variation transversale. Une théorie fort simple est indiquée pour le cône voulu de champ magnétique. Ce dernier est comparé à la pente de champ obtenu du champ de frange entre les pôles d'un aimant. Un rapport de perte inverse/avant de 80 a été réalisé expérimentalement dans la gamme de fréquence de 7,5GMHz à 11,5GMHz avec une dalle de ferrite au magnésium-manganèse F5X mesurant 0,300" × 0,100" × 3,25" et montée dans un guide d'onde standard de bande X.*

**La synthèse des générateurs de codes cycliques** par P. E. K. Chow et A. C. Davies

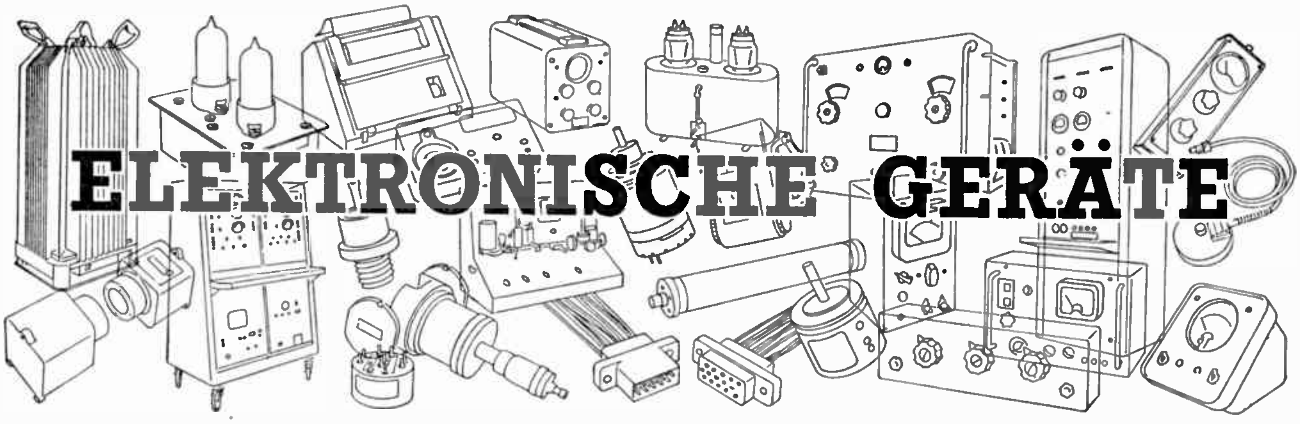
Résumé de l'article aux pages 253 à 259

*Dans la division de fréquence numérique et la production de codes, les registres de glissement à réaction comportent un certain nombre d'avantages par rapport aux méthodes plus connues utilisant les compteurs binaires. Les méthodes de réalisation de registres de glissement à réaction sont expliquées dans cet article et une nouvelle méthode est proposée. Cette méthode conduit à la réalisation d'un circuit semblable mais ayant des caractéristiques plus générales que celles du registre de glissement à réaction. Elle peut ainsi s'appliquer à un domaine d'utilisation plus étendu et elle n'est pas limitée à la production de codes en chaîne.*

**Appareil d'essai pour tubes stabilisateurs** par D. T. Smith

Résumé de l'article aux page 260

*Cet article décrit un appareil d'essai pour tubes de référence et de stabilisation de tension et pour diodes Zener. On fait passer les courants dans la gamme de 0,5 à 20mA à travers le dispositif soumis à l'essai et la tension développée est mesurée sur un voltmètre.*



# ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, **Zubehörteile** und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

*Übersetzung der Seiten 268 bis 273*

### Hall-Messkopfbausatz

**Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3**

(*Abbildung Seite 268*)

Ausnutzung des Hall-Effekts in diesem Bausatz schafft ein einfaches und bequemes Mittel zum Messen magnetischer Feldstärken. Bei Anschluss an ein 50-$\mu$A-Messgerät können Feldstärken von 100...20 000 Gauss gemessen werden. Der Bereich kann durch einen in der Bauanweisung beschriebenen einstufigen Transistor-Verstärker nach unten erweitert werden.

Die mitgelieferte Bauanweisung für den Hall-Messkopf enthält leicht verfolgbare Zeichnungen, die jeden der beim Zusammenbau zu beachtenden Schritte illustrieren, und gibt eine klare und eingehende Beschreibung der unterliegenden theoretischen Prinzipien. Der Student ist somit in der Lage, zahlreiche Hall-Effekt-Versuche auszuführen und dessen praktische Bedeutung schätzen zu lernen.

Der Hall-Messkopf kann mittels eines mitgelieferten Standard-Magneten geeicht werden. Andererseits kann man den Magnet zusammen mit dem Messkopf dazu verwenden, den Hall-Effekt vorzuführen oder Hall-Parameter zu messen.

Die Ausgangsspannung ist 200 mV für ein 10-kG-Feld und die Temperaturkonstanz 0,3 %/° C.

**EE 68 751** *für weitere Einzelheiten*

### Verzögerungs-Schnappschalter

**Standard Telephones & Cables Ltd., Connaught House, 63 Aldwych, London, W.C.2**

(*Abbildung Seite 268*)

Ein neuer Schnappschalter mit thermischer Verzögerung Type S45C/1D wird nunmehr von Standard Telephones & Cables Ltd angeboten. Der für verhältnismässig hohe Leistungen (bis zu 250 V~, 5 A) ausgelegte neue Schalter ist ein einpoliger Umschalter mit hohem Kontaktdruck und sehr schnellem Schliessen und Unterbrechen.

Die nominelle Schaltverzögerung ist 42 s, und durch Reihenschaltung einer Anzahl von Schaltern kann man ein Mehrfaches der Verzögerungszeit erlangen.

Thermische Verzögerungsschalter sind für die Verzögerung sehr geeignet, die bei indirekt geheizten Röhren zwischen der Beaufschlagung mit der Heizspannung und dem Anlegen anderer Schaltungsspannungen wie z.B. der Anodenspannung erforderlich ist. Ausserdem können sie für Kippgeneratoren sehr niedriger Frequenz, die Sterndreieckumschaltung von Drehstrom beim Anlaufen von Asynchronmotoren und das Wiederschliessen eines Leistungsschalters nach Auslösen durch einen vorübergehenden Stromstoss eingesetzt werden.

**EE 68 752** *für weitere Einzelheiten*

### Präzisions-Intervallzeitgeber

**Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex**

(*Abbildung Seite 268*)

Dieser Intervallzeitgeber wurde speziell zum Messen der Arbeitszeit von Relaiskontakten in Millisekunden entwickelt, kann aber genau so gut für die Bestimmung irgendwelcher durch einen externen Kontakt oder Kontakte begrenzten Intervalle verwendet werden.

Interne, von einem kristallgesteuerten Oszillator (Unsicherheit 0,01 %) mit 1 kHz abgegebene Impulse werden von vier Dekatron-Röhren in Kaskade während der Zeitspanne gezählt, in der externe Kontakt(e) betätigt sind, und die Endzeit (bis zu einem Maximum von 9,999 s) direkt von diesen Dekatron-Röhren abgelesen.

Mittels eines sechsstufigen Drehschalters und zweier federnder Eingangsklemmenpaare kann der Zeitgeber durch jede Kombination von ein oder zwei Kontakten, die entweder schliessen oder unterbrechen, gestartet oder gestoppt werden. Ein dreistufiger Drehschalter ist für "Rückstellen-Betrieb" vorgesehen; das Gerät ist für 200...250 V~ Netzanschluss ausgelegt.

Verschiedene Ausführungen sind mit Höchstzählung bis zu 999,99 s (in hundertstel Sekunden) lieferbar, und alle Modelle sind mit oder ohne polierten Holzkasten erhältlich.

Gegen kleinen Aufschlag können auch Ausführungen für Betrieb mit externen Impulsen oder Gleichspannungen geliefert werden.

**EE 68 753** *für weitere Einzelheiten*



### Brücken-Selbstabgleich

**The Wayne Kerr Laboratories Ltd, New Malden, Surrey**

(*Abbildung Seite 268*)

Ein von Wayne Kerr Laboratories Ltd entwickeltes neues Gerät ermöglicht genaues Messen der Wirk- und Blindkomponenten einer unbekannten Impedanz ohne Handabgleich. Der neue Zusatz "Autobalance" Type AA221 hält —wenn mit der Kerr Universalmessbrücke B221 zusammen betrieben—elektronisch einen dauernden abgeglichenen Zustand aufrecht und zeigt Änderung im Wert jeder Komponente sofort auf einem Zwillings-Messgerät an.

Die beiden Messgeräte des "Autobalance"-Zusatzes ersetzen tatsächlich die Feineinstellung an der Messbrücke. Bei geeigneter Bereichwahl kann daher entweder das ganze Messergebnis oder—bei Einstellung der ersten beiden Ziffern mittels der Dekadenschalter der Brücke —ein Gesamtergebnis mit vier geltenden Ziffern angezeigt werden. In beiden Fällen werden die Feineinstellungen auf Null gestellt und der Abgleich elektronisch bewahrt. Die Grundgenauigkeit von 0,1 %, der ungewöhnliche Messumfang und das Unterscheidungsvermögen der Brücke B221 bleiben durch Benutzung des Autobalance-Zusatzes unbeeinflusst.

Für Reihenprüfen können die Feineinstellungen der Brücke so eingeregelt werden, dass die Skalenmitte der Anzeige dem gewünschten Nennwert entspricht. Bei Verwendung einer einfachen Maske für die Skalen können auch ungelernte Kräfte Bauelemente prüfen. Die Ausgänge der Messgerätschaltungen können auch zu Buchsen hinten am Autobalance-Zusatz herausgebracht werden, um einen automatischen Partienzähler zu betätigen. Ferner kann man Digitalvoltmeter und Druckwerke, Registriergeräte und Fernanzeigen anschliessen.

Der Autobalance-Zusatz gibt sowohl Signalquelle wie auch Detektor, ist transistorisiert und aus einer internen Batterie gespeist. Die Kombination Brücke und Zusatz kann daher benutzt werden, wo Wechselstrom nicht vorhanden ist. Für Dauerbetrieb ist ein Netzgerät lieferbar.

**EE 68 754** *für weitere Einzelheiten*

### Tonfrequenzgangschreiber

**Vertrieb: Claude Lyons Ltd, Hoddesdon, Hertfordshire**

*(Abbildung Seite 269)*

Der von Construzione Elettriche di D. Borsini in Ancona, Italien hergestellte Automatische Frequenzgangschreiber TAR/61 ist ein Tonfrequenz-Spektrumanalysator. der Versuchsarbeiten und Tests an Verstärkern, Filtern, Aufnahmegeräten, Mikrofonen. Tonabnehmerköpfen, Lautsprechern und Wandlern sowie Schaltungselementen im allgemeinen, erleichtert.

Das Gerät besteht im wesentlichen aus einem Schwebungsoszillator mit geringem Klirrfaktor, dessen Frequenzbereich 0...20 kHz mittels eines Motorantriebs wiederholt überstrichen wird. Der bei variabler Frequenz konstante Ausgangspegel des Schwebungsoszillators wird über einen Leistungsverstärker mit geeichten Abschwächern dem zu prüfenden System aufgedrückt; das vom letzteren abgegebene Ausgangssignal wird zum Frequenzgangschreiber zurückgeleitet und an einen logarithmischen Verstärker gelegt, der das vertikale Ablenksystem einer Oszillografenröhre mit langer Nachleuchtdauer speist. Die Zeitablenkung der Oszillografenröhre wird einem 180stufigen Potentiometer entnommen, das auf die Durchlauftreibwelle des Schwebungsoszillators montiert ist, wodurch eine direkte Amplituden-Frequenzkurve erzeugt wird.

Der Schirmraster ist senkrecht in dB und mV, waagerecht in Frequenz geeicht, und die Spur leuchtet lange genug nach, um aufeinanderfolgende Durchläufe zum Vergleich zu überlagern, so dass das Ergebnis jeder Nachregelung sofort augenscheinlich wird. Bezugsfrequenzgänge oder Eichgrenzwerte können auf dem Raster markiert und Fertigungsprüfungen dadurch wesentlich vereinfacht werden.

Wo Zweipolsysteme wie z.B. Mikrofone oder Lautsprecher untersucht werden sollen, muss natürlich ein entsprechender Wandler verwendet werden, um das System zu komplettieren—z.B. ein Lautsprecher zum Erregen des Mikrofons. Der sekundäre Wandler kann Fehler in das Ergebnis einführen; man kann einen durch Hilfswandler gespeisten Verstärker mit Dynamikregler zur Korrektur auftretender Linearität in den Oszillatorkreis schalten.

**EE 68 755** *für weitere Einzelheiten*

### Beanspruchungs- und Temperaturerfassung

**Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex**

*(Abbildung Seite 269)*

Dieses aus zwei Geräten bestehende transportable Datenerfassungssystem wurde im Auftrag des britischen Amts für Wissenschaftliche und Industrielle Forschung entwickelt und ist für die Erfassung von Dehnungs- und Temperaturmessungen in Spannbetonkonstruktionen bestimmt.

Bis zu 50 Kanäle jedes Parameters lassen sich messen und die Ergebnisse in digitaler Form sowohl ausdrucken als auch in Lochstreifen stanzen.

Beanspruchung wird mit Hilfe von Saiten-Spannungsmessern bestimmt, die in den Beton eingebettet werden. Die Zeit für 100 volle Schwingungen wird registriert, und die einfache quadratische Beziehung zwischen Beanspruchung und gemessener Zeit erlaubt direkte Berechnung der internen Beanspruchung in der Struktur.

Für die Temperaturmessungen werden Widerstandsthermometer verwendet und die Ergebnisse als direkte Anzeige in der absoluten Temperaturskala dargestellt.

Die Ausrüstung ist für unbemannten Betrieb an Ort und Stelle ausgelegt und kann für automatischen Ablauf in vorgewählten Intervallen von bis zu 2 Stunden programmiert werden.

**EE 68 756** *für weitere Einzelheiten*

### Metallfilm-Widerstände

**Dubilier Condenser Co. (1925) Ltd, Ducon Works, Victoria Road, North Acton, London, W.3**

*(Abbildung Seite 269)*

Metallfilm-Widerstände Typ ME sind für fünf Nennbelastungen und vorbehaltlich der Bereichgrenzen der Grössenbelastbarkeit in Widerstandswerten von 30 Ω...10 MΩ lieferbar. Die normale Toleranz ist ±1 %, eingeengte Toleranzen von 0,5 %. 0.25 % und 0,1 % können aber abhängig vom Temperaturkoeffizienten geliefert werden.

Die äussere Ummantlung für die niedrigeren Leistungen besteht aus feuchtigkeitsfester Hochtemperatur-Kunststoffisolierung und für die 1-W- und 2-W-Grössen aus einer Porzellanröhre mit hermetisch abgeschlossenen Enden. Alle fünf Grössen sind unabhängig vom Widerstandswert sehr rauscharm, üblicherweise unter 0,1 μV/V.

Der zwischen einem Zehntel und voller Nennspannung gemessene Spannungskoeffizient ist für alle Werte niedriger als $5 \times 10^{-6}$/V; für niedrige und mittlere Widerstandswerte ist er praktisch Null. Vorhandene Daten zeigen, dass ein Jahr Lagerung bei normaler Raumtemperatur keine mit Standard-Laborgeräten messbaren Änderungen hervorruft. Widerstände können mit fünf Temperaturkoeffizienten von $\pm 150 \times 10^{-6}$ bis zu dem engen Koeffizienten $\pm 15 \times 10^{-6}/°$ C herunter geliefert werden.

Mit Ausnahme der niedrigen Widerstandswerte der Typen für hohe Belastbarkeit sind die Widerstände bis zu rund 100 MHz induktionsfrei. Wo es auf Miniaturisierung ankommt, können sie vor allem Präzisions-Drahtwiderstände gut ersetzen.

**EE 68 757** *für weitere Einzelheiten*

### Schwebungsoszillator

**Vertrieb: Livingston Laboratories Ltd, 31 Camden Road, London, N.W.1**

*(Abbildung Seite 269)*

Ein die gesamten Hör- und Unterhörfrequenzbänder in einem Bereich überstreichender Oszillator wird von Livingston Laboratories Ltd angeboten, und zwar der von Radiometer in Dänemark hergestellte Schwebungsoszillator HO32.

Das Gerät hat einen Gesamtfrequenzbereich von 0...21 kHz, wobei 0...20,5 kHz in einem Durchlauf überstrichen wird. Die Ausgangs-Höchstleistung ist 4 W und kann einem Ausgangsübertrager oder dem geeichten Abschwächer entnommen werden. Der Übertrager hat Abgriffe zum Anpassen der verschiedenen Lasten und die Wahl eines 15-Ω-Ausgangs. Der Abschwächer gibt Ausgangspegel von 300 μV ... 100 V ab; der Klirrfaktor liegt unter 0,1 % für Abschwächer-Ausgangsspannungen unter 10 V. Frequenzkonstanz und Frequenzgang sind gut; zur Erleichterung der Nullschwebungs-Bestimmung ist ein Abstimmauge vorgesehen.

**EE 68 758** *für weitere Einzelheiten*

### Isolationsmesser

**Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex**

*(Abbildung Seite 270)*

Der Isolationsmesser 190 ist für direktes Messen des Isolationswiderstands und Kriechstroms ausgelegt. Er ist tragbar und mit einem Transistor-Wechselrichter zur Erzeugung der Prüfspannung von einer internen Batterie ausgestattet. Prüfspannungen von 500 V, 250 V und 100 V sind vorhanden; das Messgerät hat 89 mm Skalenlänge und ermöglicht dadurch Messen von Isolationswiderständen von 1 GΩ.

Drucktastenschalten erlaubt einhändigen Betrieb und verhindert unnötige Entnahme von Batteriestrom. Ein einziger Drehschalter wird zur Einstellung der Prüfspannung betnutzt und hat eine Position "Eichung" für die Einstellung der Ausgangsspannung mittels Trimmerregler. Verwendung einer Quecksilberbatterie als kompakte Stromquelle gibt eine Lebensdauererwartung von über 25 000 durchschnittlichen Messungen bei grosser Lagerfähigkeit und Korrosionsfreiheit.

Der Ausgang wird elektronisch konstantgehalten, und der Höchststrom ist auf 50 µA begrenzt. Die Strombegrenzung reduziert die Prüfspannung proportional von ihrem vollen Wert bei offnem Stromkreis zu Null bei Vollausschlag. Durch diese Eigenschaft wird das Risiko eines zerstörenden Durchschlags beseitigt, und das Instrument kann ohne Gefahr zum Messen von Halbleiter-Gleichrichtern, -dioden usw. verwendet werden.

**EE 68 759** *für weitere Einzelheiten*

## Druck-Messwertwandler

**Intersonde Ltd, The Forum, High Street, Edgware, Middlesex**

*(Abbildung Seite 270)*

Für das Messen statischer oder dynamischer, gasförmiger oder flüssiger Drücke hat Intersonde Ltd Druck-Messwandler mit aufgeklebten Dehnstreifen mit Druckbereichen 0...70 kg/cm² bis zu 0...700 kg/cm² unter den Typenbezeichnungen PR14 und PR15 eingeführt.

Das beiden Typen gemeinsame druckempfindliche Element ist ein genaubearbeitetes Röhrchen, auf das Dehnstreifenelemente bifilar gewickelt und in eine vierzweigige Brücke geschaltet sind. Beide Messwertwandler haben einen Ausgangswiderstand von 350 Ω und erzeugen bei Erregung mit 15 V eine Ausgangsspannung von 22 mV bei vollem Nenndruck. Der kombinierte Fehler der Nichtlinearität und Hysterese ist bei Vollausschlag innerhalb 0,25% und der Betriebstemperaturbereich −40° C ... +120° C.

Typ PR14 hat Stiftanschlüsse, Typ PR15 freie Zuleitungen, und beide sind 125 mm lang und haben 25,7 mm Durchmesser mit 1/4-Zoll-Rohrgewinde für den Druckstutzen.

**EE 68 760** *für weitere Einzelheiten*

## Abstimmbares Impulsmagnetron

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

*(Abbildung Seite 270)*

Typ 7008 ist ein abstimmbares Magnetron, dessen elektrische Kenndaten denen der Serie 4J50A ähnlich sind. Es hat eine typische Höchstspitzenleistung von 220 kW und einen Frequenzbereich von 8,5...9,6 GHz. Das Tastverhältnis ist 0,001 bei 0,2...2,75 µs Impulsdauer und einem Spitzenanodenstrom von 27,5 A.

Sonderkennzeichen des 7008 sind u.a.: Mindest-Spitzenausgangsleistung 200 kW; lineare Abstimmcharakteristik mit geeichter Grob- und Feinabstimmung; die kompakte Abstimmvorrichtung ist besonders für geringes Spiel konstruiert. Kein Frequenzspringen, Fehlimpulsrate unter 0,2%, schneller Spannungsanstieg, besonders langlebige Katode, wenig oder gar kein Funken, bei gewissen Umgebungs- und Betriebszuständen Kühlung durch Strahlung.

**EE 68 761** *für weitere Einzelheiten*

## Peltier-Batterien

**Mullard Ltd, Mullard House, Torrington Place, London, W.C.1**

*(Abbildung Seite 270)*

Mullard Ltd hat für einen breiten Kühlanwendungsbereich, der eine kompakte "Kühlquelle" erfordert, drei "Peltier-Batterien" eingeführt.

Bei begrenztem Raum bieten die Peltier-Batterien bedeutende Vorteile gegenüber den herkömmlichen Kühlapparaten, da sie weder Verdichter noch Heizeinheit benötigen. Typische Anwendungsbeispiele sind das Kühlen in kleinerem Massstab in medizinischer und biologischer Forschung sowie Wärmeabfuhr in Transistor-Geräten.

In den Batterien wird Wismuttellurid als thermoelektrisches Element benutzt. Der Peltier-Effekt, den die Batterien ausnutzen, entsteht, wenn ein Gleichstrom durch die Trennebene von zwei ungleichen Metallen oder Halbleitern fliesst und besteht aus einem Temperaturgefälle an der Trennebene.

Die Batterie PT11/20 ist für einen Betriebsstrom von 18...22 A bei 1,0...1,2 V ausgelegt und hat eine Höchstkälteleistung von 16 W. Bei Dauerbetrieb mit 20 A wird eine Lebensdauer von mindestens 2 000 Stunden erwartet.

Typ PT20/20 hat einen Betriebsstrom von 20 A bei 2 V, eine Kühlungsleistung von 23 W und eine erwartete Mindestlebensdauer von 2 000 Stunden bei Dauerbetrieb mit 20 A.

Typ PT47/5 arbeitet mit 5...6 A, 5,0... 5,4 V und hat eine Kühlungsleistung von 16 W. Seine Mindestlebensdauer ist 2 000 Stunden bei Dauerbetrieb mit 5 A.

Sowohl PT11/20 als auch PT47/5 sind mit flachen Kupferplatten für Verwendung auf Festoberflächen oder mit Rippen zum Kühlen von Gasen und Flüssigkeiten lieferbar. PT20/20 hat flache Kupferplatten.

Alle drei Typen werden einsatzbereit geliefert.

**EE 68 762** *für weitere Einzelheiten*

## Koaxial-Reflektometer

**A. T. & E. (Bridgnorth) Ltd, Bridgnorth, Shropshire**

*(Abbildung Seite 271)*

Für Benutzer von UHF-Ausrüstungen hat A. T. & E. (Bridgnorth) Limited, eine Firma der Plessey-Gruppe, eine Reihe neuer Koaxial-Reflektoren für Festeinbau in Koaxialleitungen zur Überwachung auftreffender und reflektierter Leistung herausgebracht.

Merkmale sind u.a. hohe Durchgangsleistung und niedriger Einfügungsverlust. Durch parallelgeschaltete Schleifenrichtkoppler erübrigen sich Messleitungen.

Lieferbar sind Ausführungen für Frequenzbereiche zwischen 50 und 500 MHz. Das Modell für Verwendung im Band 150...300 MHz hat eine Mindestrichtwirkung von 40 dB. Die Kopplerempfindlichkeit ist 500 µA für 10 W auftreffende Leistung. Das Höchststehwellenverhältnis der primären Leitung ist besser als 1,05:1 und der Einfügungsverlust in der primären Leitung daher vernachlässigbar klein. Die Durchgangsleistung ist 300 W.

**EE 68 763** *für weitere Einzelheiten*

## Präzisions-Potentiometer

**Salford Electrical Instruments Ltd, Peel Works, Salford, 3, Lancashire**

*(Abbildung Seite 271)*

Salford Electrical Instruments Ltd hat ihr Angebot von Präzisions-Potentiometern ausgebaut, das jetzt auch Grösse 11 umfasst. Alle Abmessungen einschliesslich Drehsicherungsnase entsprechen den Vorschriften für die internationale Grösse.

Auf genaubearbeitete Ringkörper mit Chromnickeldraht oder für geräuscharme Verwendung mit Edelmetalldraht gewickelte Widerstände sind in Werten von 500 Ω...50 kΩ lieferbar.

Der Temperaturbereich ist −40° C... +85° C und die Belastbarkeit 2 W bei 20° C. Die unabhängige Linearitätstoleranz ist ±0,5%.

Die Potentiometer sind in Mehrfachausrüstung von bis zu vier gekuppelten Elementen lieferbar. Extraabgriffe können vorgesehen werden und sind jeweils an eine Einzelwindung des Drahtes angeschlossen. Goldplattierte Seitenanschlüsse sind Standardausführung zur Erleichterung des Anlötens von Verbindungen.

Sowohl für die Widerstandsbahn wie für den rhodiumplattierten Schleifring werden Doppelschleifkontakte aus Edelmetall verwendet.

Die Welle aus rostfreiem Stahl läuft in abgedeckten Kugellagern hoher Qualität.

Das Gehäuse ist aus einer hochwertigen, korrosionsbeständigen Aluminiumlegierung gearbeitet und hat einen Durchmesser von 25,55 mm.

**EE 68 764** *für weitere Einzelheiten*

## 'Prestincert'-Werkzeugausrüstungen

**Belling & Lee Ltd, Great Cambridge Road, Enfield, Middlesex**

*(Abbildung Seite 271)*

Belling & Lee Ltd hat eine neue Prestincert-Werkzeugausrüstung für Verwendung in den Sektoren Elektronik, Elektrotechnik und leichter Maschinenbau auf den Markt gebracht.

Das Prestincert-Bauelemente-System umfasst u.a. selbstlochende Gewindebuchsen und Anschlussstützpunkte für Montage auf Blech und viele Kunststoff-Isoliermaterialien in einem Arbeitsgang. Man braucht weder zu bohren, noch zu nieten, und die Befestigung ist im normalen Gebrauch praktisch unzerstörbar. Die neue Ausrüstung wurde geschaffen, um den steigenden Bedarf von Konstrukteuren und Technikern zu decken, die sich dieses Fertigungsverfahrens in der Entwicklung bedienen und das arbeitsparende System zukünftig in die Fertigung einführen wollen.

Die Ausrüstung besteht aus einer kleinen, leichten Presse mit Handbedienung, die bis zu 0,77 kg/mm² entwickelt, und sechs Behältern mit einem Sortiment von Prestincert-Bauelementen der in der Leichtindustrie gängigsten Grössen.

Für elektrotechnische und elektronische Anwendungsgebiete sind drei Gewindebuchsen mit britischen Gewinden 6BA, 4BA und 2BA sowie drei Anschlussstützpunkte in das Sortiment aufgenommen. Nachlieferung aller Typen erfolgt vom Lager.

Ausserdem werden sechs farbgekennzeichnete Einsatzwerkzeuge für die Bauelemente mitgeliefert, die mit Schiebesitz in die Presse passen und dort durch eine interne federnde Klemme festgehalten werden.

Werkzeuge, Presse und Prestincert-Bauelemente werden in einem robusten Transportkasten geliefert.

**EE 68 765** *für weitere Einzelheiten*

### Magnetköpfe

**EMI Electronics Ltd, Hayes, Middlesex**

(*Abbildung Seite 271*)

EMI Electronics Ltd hat ein neues Fertigungsprogramm für Magnetköpfe hoher Präzision—die Baureihe Emidata E—angekündigt, die speziell für Mehrspur-Messwerterfassung und -Datenverarbeitung konstruiert sind, in denen hohe Reproduzierbarkeit der Leistung erforderlich ist.

Die ungewöhnlichen Leistungskenndaten und lange Lebensdauer aller Köpfe in einem Stapel wird durch strenge Kontrolle der Spalttiefe in der Fertigung erreicht. Spalt-Ausfluchtung und Azimut der Köpfe sind viel genauer als die der Vorläufer-Baureihe S, mit der sie jedoch in jeder Beziehung austauschbar sind.

Die mittlere Azimut- und Spaltstreuung der in einer Linie liegenden Kanten der Spalte eines Stapels fällt in eine 5 μm breite Zone rechtwinklig zur Auflagefläche. Der Erhebungswinkel, bzw. die Kopfstapelneigung, wird innerhalb einer ähnlichen 12,7 mm breiten Zone gehalten. Die Standard-Spaltbreite ist 12,7 μm für Aufnahme und 5 μm für Wiedergabe.

Die Köpfe haben einen Hochfrequenzgang bis zu 100 kHz und sprechen auf eine Langwellenlänge von 2,5 mm an. Die Übersprechungsdämpfung ist äusserst gut. Die Spur der Aufnahmeköpfe ist breiter als die der Wiedergabe für den Fall, dass das Band Seitenschwankungen hat.

Köpfe für alle Spuranordnungen der I.R.I.G. und S.B.A.C. bis zu 33 gegeneinander versetzte Spuren auf 1"-Magnetband sind lieferbar; sie haben alle angeschlossene Cannon - Miniaturstecker. Köpfe für Sonderzweck-Spurabstände und Spurbreiten können gegen Auftrag hergestellt werden.

**EE 68 766** *für weitere Einzelheiten*

### Miniaturservoverstärker

**Evershed & Vignoles Ltd, Acton Lane, Chiswick, London, W.4**

(*Abbildung Seite 272*)

Der Miniatur-Transistorverstärker Evershed PA.86T ist für das Treiben eines Einphasen - Wechselstromservomotors bei Frequenzen von 50...400 Hz ausgelegt und gibt eine Höchstleistung von 10 W bei 20-0-20 V an eine Impedanz von 35 Ω je Wicklung ab.

Der Verstärker ist mit fünf Transistoren bestückt, und die Bauelemente sind innerhalb des Kühlkörpers des Endtransistors angeordnet. Komplette Umhüllung mit Silikon-Kautschuk verhindert Eindringen von Schmutz und Feuchtigkeit und erhöht den Widerstand des Verstärkers gegen mechanische Stösse. Alle Verbindungen werden zu einer Schraubklemmleiste herausgeführt.

Für Anschluss an das Wechselstromnetz ist ein Eingangsumspanner Evershed PA.43597 mit Nennausgangsspannungen von 23-0-23 V lieferbar, der drei Verstärker speisen kann. Zusätzliche Wicklungen sind vorhanden für Motor- und Generatorbezugsphasen und für den Zerhacker-Umspanner (falls erforderlich).

**EE 68 767** *für weitere Einzelheiten*

### Ablötwerkzeug

**Charles Austen Pumps Ltd, Petersham Works, High Road, Byfleet, Surrey**

(*Abbildung Seite 272*)

Das Ablöten und Entfernen fehlerhafter Bauelemente von gedruckten Schaltungen kann Schwierigkeiten bereiten, besonders wenn es sich um mehrpolige Anschlüsse handelt. Bei Benutzung eines grossen Lötkolbens besteht die Gefahr, benachbarte Bauelemente durch die Wärme derart zu beschädigen, dass die Reparatur des Bausteins unwirtschaftlich wird. Durch verlagerten Lotüberschuss werden manchmal Kurzschlüsse hervorgerufen.

Im Hinblick auf die steigende Verwendung gedruckter Schaltungen hat Charles Austen Pumps Ltd den Austen-"Soldermaster" speziell zur Überwindung dieser Schwierigkeiten entwickelt. Das Gerät ist ein kleiner Lötkolben hoher Temperatur mit hohler Kupferspitze, durch die mittels einer Pumpe und unter Kontrolle der Arbeitskraft ein Vakuum angelegt werden kann. Das durch den Kolben geschmolzene Lot der Verbindung wird durch die Spitze in ein Fanggefäss gesaugt. Auch unerwünschte Lötmitteldünste werden abgesaugt. Die Pumpe wird durch ein Filter mit porösem Kunststoffelement, das gereinigt werden kann, geschützt. Das erstarrte Lot kann nach Beendigung des Arbeitsvorgangs aus dem Fanggefäss entfernt werden.

Die Ausrüstung besteht aus einer Austen-Capex-Pumpe mit Austen-Mikrofilter, Lotfanggefäss und Rohren, 50-W-Lötkolben mit hohler Spitze und einer Ersatzspitze. Sie wird komplett mit einer Kolbenauflage und Traggriff geliefert. Abmessungen sind 190 × 152 × 152 mm hoch, Gewicht 2,7 kg.

**EE 68 768** *für weitere Einzelheiten*

### Zeit-Frequenzmesser

**Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey**

(*Abbildung Seite 272*)

Venner Electronics Ltd hat ein kompaktes Mehrzweckgerät TSA 3436 eingeführt, das für Frequenz und Periodendauer bis zu 1,2 MHz sowie für Zeitmessungen von 1 μs...100 000 s eine sechsstellige Anzeige gibt. Es kann auch bei Geschwindigkeiten bis zu 1 MHz als Zähler eingesetzt werden.

Das volltransistorisierte, tragbare Gerät TSA 3436 ist bei Abmessungen von 362 × 98 × 267 mm für Verwendung im Labor oder industrielle Qualitätskontrolle gedacht. Hauptmerkmale sind u.a. drei Durchlassschaltungen, wahlweises Austasten zwischen Ergebnissen und Selbstprüfung. Das Gerät kann Periodendauern von entweder 1 oder 10 Schwingungen in einer Wahl von sechs Zeiteinheiten messen; ein Wechselstrom-Gleichstrom-Eingangsschalter gibt höhere Empfindlichkeit für Wechselstromeingänge, während das Gerät bei Gleichstrom erst bei niedrigerer Empfindlichkeit anspricht.

Das 1-MHz-Quarz wird in einem Thermostaten auf konstanter Temperatur gehalten; die sich ergebende Frequenz wird in Binärstufen bis zu 0,1 Hz herunter geteilt.

**EE 68 769** *für weitere Einzelheiten*

## Ausbildungs-Digitalrechner

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

*(Abbildung Seite 272)*

Der "LAN-DEC"-Digitalrechner wurde für Ausbildungsanstalten entwickelt und stellt ein preisgünstiges, hochwertiges Gerät dar, mit dem man eine grosse Auswahl von Rechnerschaltungen vorführen kann.

Der "LAN-DEC" wird normalerweise als Tischgerät benutzt, kann aber auch nach Entfernung der Seiten bündig in ein 19"-Gestell eingebaut werden.

In Grunde genommen ist der LAN-DEC ein grosses Programmierfeld—483 × 267 × 645 mm—auf dem Standardsymbole für Rechnerlogik die 15 NOR - Transistor - Widerstand - Logikschaltungen darstellen, die direkt auf der Unterseite in fünf vertikalen Kolonnen von je drei Torschaltungen verdrahtet sind. Verbindungen zu jedem der NOR-Eingänge und -Ausgänge wird durch hochwertige, versilberte Miniaturbuchsen hergestellt, die für jeden Punkt paarweise vorhanden sind.

Jeder Rechner wird mit den für das Zusammenschalten erforderlichen Schnüren usw. sowie einem Handbuch geliefert, das eine grosse Auswahl von Rechnerschaltungs-Vorführungen mit Anweisungen für das Zusammenschalten sowie Zeichungen gibt. Weitere Exemplare können getrennt bestellt werden.

Ein Zusatz-Relaissteuergerät RCU.5 ist zur Steuerung von Ausseneinrichtungen lieferbar.

Im Rechner werden durchweg Transistoren OC83 verwendet, deren Fähigkeit, bis zu 0,5 A Überlastung auszuhalten, einen viel grösseren Sicherheitsfaktor als üblich gibt.

"LAN-DEC" wurde in Zusammenarbeit mit den Fakultäten der Elektrotechnik und Mathematik des College of Technology Enfield und der County Grammar School Edmonton entwickelt.

**EE 68 770** *für weitere Einzelheiten*

## Gleichstromversorgung

**De La Rue Frigistor Ltd. Canal Estate, Langley, Buckinghamshire**

*(Abbildung Seite 272)*

De La Rue Frigistor Ltd, eine Gesellschaft der De-La-Rue-Gruppe, hat eine Serie von Gleichstromversorgungen mit Netzanschluss eingeführt.

Jedes Gerät gibt einen hohen Niederspannungs-Gleichstrom ab, der von Hand kontinuierlich geregelt werden kann und so geglättet ist, dass die Restwelligkeit in jedem beliebigen Teil des Bereiches 10% nicht übersteigt.

In drei ansprechend ausgeführten Gehäusen werden elf verschiedene Geräte geliefert, die 15 oder 30 A bei 1½, 3, 6 und 9 V oder 60 A bis zu 6 V abgeben.

Jedes Standardgerät hat einen Reglerknopf, Ammeter, Signallampen und Sicherung. Die Geräte werden für Einbau in Ausrüstungen bedeutend billiger, jedoch ohne Regler mit vorgewähltem Gleichstromausgang geliefert.

**EE 68 771** *für weitere Einzelheiten*

## Peilempfänger

**Brookes & Gatehouse. Ltd. Bath Road. Lymington, Hampshire**

*(Abbildung Seite 273)*

Dieser mit "Homer" bezeichnete Peilempfänger, dessen Abmessungen nur 210 × 124 × 86 mm sind, ist für Jachten und Fischerboote bestimmt.

Der "Homer" hat folgende Einsatzmöglichkeiten:

(1) Peil-Standortbestimmung bei Einsatz mit der "Heron"-Antenne. Die überstrichenen Frequenzen umfassen alle See- und Luft-Funkfeuer der Welt (190 ... 410 kHz);

(2) Standortbestimmung mittels Konsol und Konsolan;

(3) Mit Modell K: Empfang der Wettervoraussagen, Schiffsverkehr usw. in den folgenden Bändern:
Langwellen (160 ... 250 kHz)
Mittelwellen (550 ... 1600 kHz)
Marine-Funkfernsprechband (1,6 ... 4,1 MHz).
Mit Modell L: Empfang der Wettervoraussagen in europäischen Gewässern im Langwellenband (160 ... 250 kHz). Dieses Modell hat weder Mittelwellen- noch Funkfernsprechband.

Der "Homer" ist in ein festes, hermetisch dichtes Gehäuse aus nylonüberzogener Aluminiumlegierung eingebaut. Die Batterien sind in einem getrennten Abteil auf der Rückseite des Gerätes untergebracht, das mit seinem eignen Deckel abgedichtet ist. Erforderlich sind vier Quecksilber-Batterien Mallory ZM9 (RM9 in USA), die im allgemeinen überall in der Welt zu haben sind und ungefähr 250 Stunden Dauerbetrieb geben. Ein Trägerarm für Montage an Schotte oder Instrumentenbrett wird mitgeliefert. Alle Befestigungen sind aus rostfreiem Stahl gefertigt, die Wellen der Kontrollelemente mit Kunstgummi abgedichtet und die Steckverbindungen palladiumplattiert, um Seewasser-Korrosion zu verhindern.

Eine Drahtantenne in 4,5 oder 9 m Höhe wird für Rundfunk- und Funkfernsprechempfang benutzt. Die Antenneneingangsimpedanz ist sehr niedrig, so dass in Holz- oder Kunststoffbooten stehende Verspannungen ohne Einfügen von Isolatoren als Antenne benutzt werden können. Auf Langwellen kann für Empfang von Rundfunk oder Konsol/Konsolan entweder die Drahtantenne oder die Peilantenne "Heron" eingesetzt werden.

"Heron" ist eine Richtantenne, die in Verbindung mit einem Funkempfänger Standortbestimmung durch Kreuzpeilung von Funkfeuern ermöglicht. Ein Prinzip, das auf Verwendung eines abgestimmten Ferritstabs beruht, gibt dem Gerät hervorstechende Vorteile in hoher Genauigkeit und kleinen Abmessungen.

In den Modellen A, B, C und D ist "Heron" mit einem Kompass zusammengebaut, so dass man Peilwinkel in Bezug auf den magnetischen Nordpol beobachten kann. In dieser Anordnung werden Fehler durch wandernden Kurs oder Krängung, die den herkömmlichen Peilsystemen eigen sind, vermieden. Für Stahljachten, deren Kompassabweichung nicht ausreichend korrigiert werden kann, steht ein Modell E zur Verfügung, das eine herkömmliche 360°-Azimutskala und Schwenkscheibe sowie zur Vermeidung von Krängungsfehlern Kardanaufhängung hat.

**EE 68 772** *für weitere Einzelheiten*

## Treiberverstärker für Galvanometer

**S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex**

*(Abbildung Seite 273)*

S.E. Laboratories (Engineering) Ltd kündigt einen transistorisierten Treiberverstärker S.E.425 an, der für das Treiben von HF-Galvanometern aus schwachen Quellen wie z.B. Telemetrie-Verstärkung, Tonfrequenzoszillatoren, Wiedergabediskriminatoren usw. ausgelegt ist. Weitere Anwendungsgebiete sind das Messen schwacher Signale bei Hochspannungspegeln. z.B. das Prüfen von Schaltstössen in elektrischen Anlagen, Motoranlassschaltungen usw.

Sechs elektrisch entkoppelte Verstärker auf gedruckten Schaltungen mit Steckleisten sind auf einem gemeinsamen Chassis in einem 19"-Gestelleinschub angeordnet.

Zu den hervorstechenden Eigenschaften dieses Verstärkers gehören sein hoher Rauschabstand von 1000 : 1, seine sehr niedrige Drift von weniger als ±0,5% im Temperaturbereich 0 ... 45° C und sein breiter Frequenzgang, gerade innerhalb ±2% von 0 ... 10 kHz. Nebensprechen zwischen Verstärkern ist unter ±2 $mV_{SS}$, verglichen mit der vollen Ausgangsspannung von ±3,2 V an einer 50-Ω-Last.

Die Eingangsverbindungen zu den Verstärkern sind elektrisch voneinander entkoppelt und können 250 V gegen Masse aushalten. Eingänge können entweder symmetrisch oder unsymmetrisch sein. Sechs dreipolige Einzelstecker auf der Frontplatte sind in einer Mehrfachsteckverbindung am hinteren Gestellrahmen zusammengefasst. Wenn die Verstärker mit Galvanometern zusammen verwendet werden. geben sie ausgezeichnete Gleichtaktunterdrückung.

Der Ausgangsstrom des Verstärkers ist auf ungefähr 120 mA begrenzt, um die Galvanometer gegen übermässige Überlastung zu schützen. Die Konstruktion ist—wie bei anderen elektronischen Ausrüstungen von S.E.L.—robust, die Schaltung erprobt und zuverlässig.

**EE 68 773** *für weitere Einzelheiten*

**Tragbarer Streifenschreiber**

**Leeds & Northrup Ltd, Wharfdale Road, Tyseley, Birmingham, 11**

(*Abbildung Seite 273*)

Leeds & Northrup Ltd hat eine neue Ausführung ihrer Serie Speedomax "H" entwickelt. die als "Compact A.Z.A.R." geführt wird. Es ist ein tragbarer 165-mm-Streifenschreiber, ein vielseitiges Instrument mit einem breiten Anwendungsbereich für Forschung und Prüffeld.

Er hat getrennte und unabhängig von Quecksilber-Batterien gespeiste Null- und Umfangregelung und kann jeden beliebigen geeichten Spannungsumfang zwischen 0.3 und 50 mV und jede beliebige Nulllunterdrückung zwischen −50 und +50 mV geben. Auf Grund dieser Eigenschaften ist es möglich, die verschiedensten Messungen aufzuzeichnen, z.B. Temperatur, Beanspruchung, Dehnung, Position, Schwingung und andere EMK-Variable einschliesslich Gaschromatografie.

Ein "Betrieb-Eichung"-Schalter ermöglicht Einstellen von Umfang und Null ohne externe Eichausrüstung. Die Messschaltungen sind für niedrigste Stromentnahme aus den Quecksilber-Batterien bemessen, um lange Lebensdauer zu gewährleisten. Ansprechen auf Umfangsregelung erfolgt nominell in einer Sekunde. Der Umfang ist mit einer Genauigkeit von unter 0,5 µV und Null mit besser als 0,0033% des Ausschlags einstellbar. Die Spannungskonstanz mit einer neuen Quecksilber-Batterie ist 0,025% des Ausschlags.

Als wahlweiser Zubehör zum Compact 'H' AZAR wird empfohlen: Vorschubgetriebe, geneigter Streifentisch für Notizen mit Abreissvorrichtung, Verstärker mit ziemlich hoher Impedanz für Nennleistung mit externer Schaltungsimpedanz von 0 ... 18 kΩ, Füsse und Traggriff.

**EE 68 774** *für weitere Einzelheiten*

# Zusammenfassung der wichtigsten Beiträge

**Ein Registrierstreifenleser und statistischer Analysator** von J. A. Phillips

Zusammenfassung des Beitrages auf Seite 216-220

*Die beschriebene Ausrüstung liest Registrierstreifen, leitet den mittleren Ausschlag, die Varianz und ein Histogramm ab. Das Abtastsystem arbeitet mit einer Vidikonröhre, ist vollautomatisch und gibt einen binärkodierten Ausgang. Für die statistischen Berechnungen wird ein kleiner Rechner mit Festprogramm eingesetzt.*

**Eine transistorisierte Schaltung für Xenon-Lichtblitzröhrenbetrieb** von P. G. M. Dawe

Zusammenfassung des Beitrages auf Seite 220-221

*Eine Schaltung wird beschrieben, in der ein transistorgesteuerter Schalter, der durch Impulse von einem Flip-flop und Multivibrator betätigt wird, eine Xenon-Lichtblitzröhre zündet. Ein Merkmal der Transistorschaltung ist die Beschaltung der Lampe mit einer Hilfsionisierungsschaltung zur Aufrechterhaltung eines Dauerentladungsweges innerhalb derselben.*

**Antennen mit Frequenzabtastung** von M. F. Radford

Zusammenfassung des Beitrages auf Seite 222-226

*Eine Antenne mit Frequenzabtastung ist eine Antenne, die so konstruiert ist, dass die Richtung des ausgesandten Strahls eine Funktion der Frequenz ist. In diesem Beitrag werden die allgemeinen Prinzipien solcher Antennen erläutert, und es wird gezeigt, dass sie eine wirtschaftliche Alternative zu mechanisch bewegten Antennen sein könnten.*

**Einige einfache Dekatron-Ankoppelschaltungen** von A. J. Oxley

*Zusammenfassung des Beitrages auf Seite 227-232*

*Vier Ankoppelschaltungen mit Serien- und Parallelübertragung, von denen drei für umkehrbares Rechnen geeignet sind, werden beschrieben. Die Höchstzählgeschwindigkeiten werden für jede Schaltung angegeben; sie sind alle niedriger als die durch die Dekatron-Röhre gesetzte Begrenzung. Wo das annehmbar ist, können—wie gezeigt wird—sehr einfache Ankoppelschaltungen verwendet werden.*

**Ein neuer frei laufender Multivibrator** von V. M. Ristic

*Zusammenfassung des Beitrages auf Seite 232-233*

*Der Beitrag beschreibt einen seriengeschalteten, emittergekoppelten Multivibrator mit zwei pnp-Transistoren. Die Schaltung hat gute Anstieg- und Abfallzeiten und schwingt zwischen 0,01 Hz und 250 kHz.*

**Eine neue Methode zur Erkennung von ohmschem Widerstand in Gegenwart von Blindwiderstand** von E. R. Wigan

*Zusammenfassung des Beitrages auf Seite 238-239*

*In diesem Beitrag wird eine Brückenmethode beschrieben, nach der ohmscher Widerstand in Gegenwart von Blindwiderstand erkannt und auf Wunsch gemessen werden kann. Die theoretische Grundlage für dieses Messverfahren wird eingehend besprochen, jedoch werden keine genauen Schaltungsanordnungen beschrieben.*

**Ein Transistor-Verstärker mit Gleich- und Wechselstrom-Gegenkopplungsstabilisierung** von H. C. Bertoya

*Zusammenfassung des Beitrages auf Seite 240-245*

*Der Beitrag beschreibt und analysiert einen Transistorverstärker, der zum Stabilisieren des Gleichstrom-Arbeitspunktes einen Gleichstrom-Gegenkopplungsweg benutzt. Sollte der gewünschte Wechselstrom-Gegenkopplungsfaktor nicht mit dem der Gleichstrom-Gegenkopplung zusammenfallen, besteht die Möglichkeit, zwei getrennte Gegenkopplungswege vorzusehen.*

*Zwischen den Stufen wird direkte Kopplung angewendet, woraus sich ein einfacher und wirtschaftlicher Entwurf ergibt.*

*Praktische Beispiele werden sowohl für pnp- als auch npn-Transistoren gegeben. Eins der Beispiele wurde speziell für Anwendungen, die rauscharme Verstärker erfordern, entwickelt.*

**Ein verbessertes Verfahren zum Messen des Hautwiderstands** von L. E. H. Lindahl, A. H. Christianson und R. N. Gooch

*Zusammenfassung des Beitrages auf Seite 245-246*

*Ein Verfahren zum Messen des Hautwiderstands wird beschrieben, in dem die Substitutionsmethode angewandt wird. Der benutzte symmetrische Gleichstromverstärker überwindet die Nichtlinearität, niedrige Empfindlichkeit und unzureichende Ausgangsleistung vieler bestehender Geräte. Erfahrungen haben bewiesen, dass die Konstanz beträchtlich innerhalb der für psychophysiologische Arbeiten erforderlichen Toleranz liegt.*

**Silizium-Lawinengleichrichter** von G. Duddridge

*Zusammenfassung des Beitrages auf Seite 247-249*

*Die Herstellung und Eigenschaften der Silizium-Lawinengleichrichter werden beschrieben, und die Fähigkeit dieser Gleichrichtertypen, Überspannungsstösse in Sperrichtung aufzufangen, wird hervorgehoben. Sowohl Serien- wie Parallelbetrieb werden behandelt und Anwendungsmöglichkeiten beschrieben.*

**Ein Breitband-Resonanzisolator mit konischem Feld** von J. H. Collins und T. M. Heng

*Zusammenfassung des Beitrages auf Seite 250-252*

*Der Beitrag beschreibt Entwicklungsarbeiten an dem Entwurf eines Breitband-Resonanzisolators mit einem querveränderlichen externen Gleichstrom-Magnetfeld. Eine einfache Theorie für die erforderliche Verjüngung des Magnetfeldes wird gegeben und ein Vergleich mit dem Feldgefälle, das man vom Randfeld zwischen den Polen eines Magneten erhält, angestellt. In Versuchen wurde mit einem Magnesium-Mangan-Ferritkern mit Abmessungen von 0,3″ × 0,1″ × 3,25″ (7,6 × 2,54 × 82,55 mm) in einem Hohlleiter für das X-band und über den Frequenzbereich 7,5 . . . 11,5 GHz ein Sperr- zu Durchlassdämpfungsverhältnis von 80 erzielt.*

**Die Synthese zyklischer Verschlüsselungen** von P. E. K. Chow und A. C. Davies

*Zusammenfassung des Beitrages auf Seite 253-259*

*Rückführungs-Verschieberegister haben für die digitale Frequenzteilung und Codeerzeugung gewisse Vorteile gegenüber den bekannteren Verfahren mit binären Zählern. Methoden für den Entwurf von Rückführungs-Verschieberegistern werden erklärt und eine neue vorgeschlagen. Diese Methode führt zu einer ähnlichen, aber allgemeineren Schaltung als die des Rückführungs-Verschieberegisters, die einen breiteren Anwendungsbereich hat und nicht auf die Erzeugung von Kettencodes beschränkt ist.*

**Glimmstrecken-Tester** von D. T. Smith

*Zusammenfassung des Beitrages auf Seite 260*

*Dieser Beitrag beschreibt einen Tester für Glimmlampenstabilisatoren und Bezugsröhren, sowie Zenerdioden. Ströme im Bereich 0,5 . . . 20 mA fliessen durch die zu prüfenden Röhren, und die auftretende Spannung wird an einem Voltmeter gemessen.*