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Commentary

COMPUTERS and aeronautics have both been very much in the news during the last few weeks and both have been the subject of urgent talks at Government level. On the one hand there has been an exhortation to buy British, while on the other there has been, so to speak, a threat to buy from abroad.

Computers came into the news when, at the end of December, Mr. George Brown, Minister of Economic Affairs, told leading industrialists and trade unionists that one of the most urgent tasks facing the country was to improve exports of electronic equipment and that the Government is worried by the large scale of imports of such products as computers, the bulk of which comes from the United States of America. The cause for this concern can be seen from the appropriate figures: in 1963 Britain exported electronic computers worth £3 779 789 against imports of £8 401 384; for the first ten months of 1964 exports were valued at £3 816 139 while imports were £9 889 926.

A detailed study has already been started on the 'extent and causes' of imports and this is being undertaken by the electronics NEDC, on which both employers and trades unions are represented. It has been stated that Mr. Brown intends to be present at future meetings of this 'Little Neddy' as he regards the electronic industry as being of such vital importance.

While many people in the electronic industry, and in industry and commerce generally, will have serious doubts as to the competence of the present committee to deal with this problem, there can be no doubt but that the problem exists and it is made all the more sombre and depressing when it is recalled that only fifteen years ago Britain had an undoubted lead in the field of computers.

The position was stated quite fairly by Professor Stanley Gill, who occupies the new Chair of Computer Sciences in the Electrical Engineering Department of Imperial College, when he gave a lecture at a recent meeting of the Association for Science Education. He stated that "Whereas in 1950 we were leading the world in this subject we have completely and utterly failed to rise to the occasion. We are now even falling behind the rest of the Western World in the rate we are developing and exploiting computers. The reason is that we have our technological scales of value all wrong. We look for the obvious immediate successes, like the bomb with a bigger bang or an aeroplane that goes that little bit faster . . . The importance of this technology in the future is, overall, greater than that of atomic energy or space travel. It is the really fundamental technology that is going to lie behind and support, and cause great changes in, the other technologies we are developing today."

The position which Britain occupied in the computing field only a comparatively short time ago is highlighted by the recent announcement that the world's oldest operating computer, LEO I, which first ran in February 1951 and was demonstrated to the Queen (then Princess Elizabeth) at that time, finally closed down in the first week of January after more than thirteen years of continuous operation. This machine was the first electronic computer in the world to go into commercial operation and was used by J. Lyons & Co. Ltd for the vast majority of their clerical operations—costings, invoicing, ordering, stock control and a payroll for 20 000 employees. It was fully described in a series of articles in this journal. In 1953 it worked out the first P.A.Y.E. income tax tables to be calculated on a computer for the Inland Revenue. LEO I also produced the actuarial mortality tables used by all insurance companies and processed a vast amount of scientific work for Government departments, aircraft companies and universities. It inaugurated the first commercial computer service bureau in the world, and among the earliest customers were the Ford Motor Co. Ltd, and the Ever Ready Co. Ltd.

The present case of the computer industry is, unfortunately, by no means unique, for over the years there have been far too many instances where Britain has pioneered a development only to see its commercial exploitation carried out elsewhere, often, of course, in America. One of the basic causes for this situation is the fact that we have a comparatively small home market to support the very large development costs of these highly technological projects. This is probably aggravated to some extent by the reluctance of British industrialists to depart from traditional methods and to investigate the uses of new tools such as computers. In the case of computers, another contributory factor to the present position may well lie in the true ownership of the firms concerned: the American investment in the industry is undoubtedly large, and to such firms it is probably of little concern whether the equipment is manufactured in this country or whether parts are imported for assembly.

It is obviously necessary that this unsatisfactory situation should be subjected to detailed study and it is imperative for the wellbeing of the country as a whole that a cure should be found for these ills.

So far as the aeronautic industry is concerned it is as yet too early to comment, for the exact intentions of the Government are not yet known. The aviation section of the electronic industry is an important one and any diminution of the prestige of the British aircraft industry may well have wide repercussions on it. It may well prove to be considerably harder to sell British radar, navigational and automatic landing systems if we are not at the same time in the van of aircraft design and production. These subsidiary factors are no doubt well known to the Government and it can only be hoped that it will give due weight to them when making its decisions.

Fading Machine for the Simulation of the Ionosphere

By D. Beck*, B.Sc., and J. A. Betts*, Ph.D., A.M.I.E.E.

The fading machine simulates the random variations of amplitude and phase angle of a long distance h.f. radio signal by addition of the outputs from eight crystal oscillators nominally at 100kc/s. Allowance is made for simultaneous propagation over two paths with a path time difference of up to 2msec. In its present form the machine permits single aerial working only but a duplication of the equipment would provide dual diversity reception. White Gaussian noise may be added to the signal to provide signal-to-noise power ratios from 0 to 50dB. The input, in the range 300 to 6 000c/s may be either a radiotelegraph or any single sideband signal.

(Voir page 136 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 143)

LONG distance h.f. radio systems experience fading and multi-path propagation due to the characteristics of the ionosphere. The multi-path effect produces frequency selective fading and it is now well known that this may be used to advantage in radiotelegraph systems[1].

Practical tests to compare various sytems are expensive and a method of simulating the ionosphere under controlled conditions is desirable for laboratory work. The fading machine described below is based on the equipment developed by H. B. Law at the Post Office Research Station[2]. The present design is transistorized and will form a transportable unit with its associated error detecting and counting equipment. It is particularly suitable for multichannel radiotelegraphy development but may be used with any s.s.b. radiotelephone system having a bandwidth of 300 to 6 000c/s. A simple modification would also permit d.s.b. signals to be investigated. For convenience the output signal is restored to the original frequency range 300 to 6 000c/s. Modifications of the local oscillator frequency to the output demodulator would, however, allow any desired frequency band to be selected. The input impedance of the fading machine is 600Ω and the minimum required signal is 1mW.

Characteristics of a Fading Signal due to Ionospheric Propagation

It has been shown that the random amplitude variations of the signal conform closely to the Rayleigh probability distribution[3],

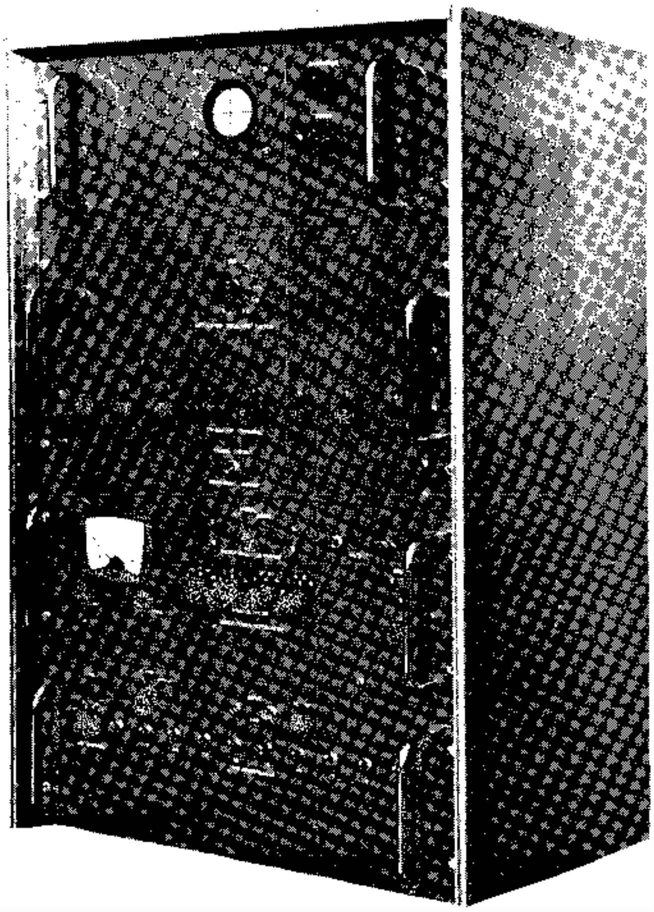
$$P(p) = 1 - \exp(-(p/S))$$

where S is the mean signal power and $P(p)$ is the proportion of the time for which the signal power is less than p. The Rayleigh distribution is obtained when a large number of equal amplitude signals, having random phases but nominally the same frequency, are added together[4].

The speed of fading is usually defined as the frequency at which the signal amplitude increases through a level corresponding to half-mean signal power. Observations have shown[2] that this quasi fading period for long distance h.f. radio signals is of the order of 4 to 15 fades/min.

Propagation over two or more simultaneous paths produces a resultant signal which is dependent upon the path time differences. Previous work by J. W. Allnatt and E. D. J. Jones of the Post Office Research Station has shown that path time differences of up to 2msec may be experienced.

The complete equipment



Basic Design Requirements

The simulation of a carrier signal with Rayleigh fading requires the combination of a large number of random phase carrier components. A good approximation is realized when the number of components is equal to or greater than six[2]. The fading rate is dependent upon the frequency spread of the carrier components. With uniform spacing of the component frequencies, the quasi frequency of fading is approximately half the carrier frequency spread. For example, a fading rate of 10 fades/min requires a frequency spread of 0·33c/s and therefore the frequency spacing between each of the carrier components would be approximately $(0{\cdot}33/n - 1)$c/s, where n is the number of carrier components. The stability of the crystal oscillators generating the components must allow this spacing to be maintained over a typical test

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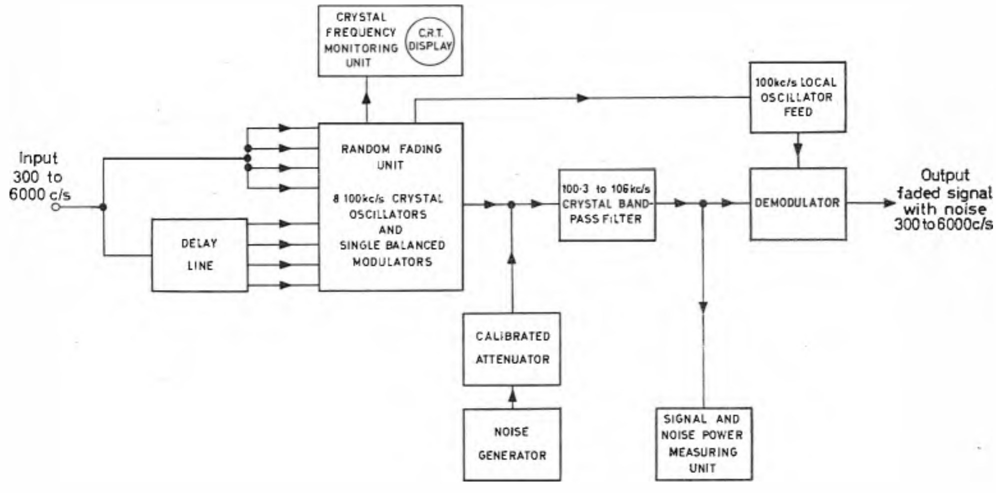


Fig. 1. Arrangement of fading machine

period of 15min. If two adjacent carrier components produce a zero beat, due to oscillator drift, then the effective number of components forming the resultant fading signal is reduced by one. This cannot be offset by using a large number of components since the required frequency spacing becomes very small and greater stability of the crystal oscillators would be required. A compromise must be made between achieving a good approximation to the Rayleigh signal and obtaining the necessary oscillator stability with inexpensive crystals and simple circuits. For this reason eight carrier components are used.

Oscillator stability tests over a 15min period indicate that a frequency spacing of 0·05c/s between carrier components can be achieved without difficulty. This limits the minimum fading rate to 10 fades/min. This should not be a serious disadvantage since it is the rapid fading rates which adversely affect the performance of radio links.

The carrier frequency of the simulated fading signal need not be in the range 3 to 30Mc/s, i.e. the h.f. spectrum, since any desired ionospheric characteristics may be selected and superimposed on the carrier. From practical considerations of filter design and crystal oscillator stability, a 100kc/s carrier frequency was chosen.

Circuit Description

The input signal is fed through a resistive splitting network to eight balanced modulators each of which is supplied with a carrier component nominally at 100kc/s. Multi-path propagation is simulated by inserting a time delay between the input signal and four of the balanced modulators, see Fig. 1. The eight modulated outputs are added together to form the fading signal.

The fading rate is determined by the frequency spacing of the carrier components. The frequencies of the crystal oscillators may be varied within small limits and the frequency difference between any two adjacent components measured on the Frequency Spacing Monitor. The fading signal is passed through a 100·3 to 106kc/s crystal band-pass filter to remove the lower sideband and residual carrier. White noise with uniform spectrum density over the range 100 to 106kc/s can be inserted at the input of the crystal filter. The calibrated attenuator and thermocouple power measuring unit provide for the adjustment of signal-to-noise ratio.

The output from the band-pass filter is demodulated with one of the 100kc/s carrier components to restore the signal to the original input frequency.

Input Circuit and Delay Line

The 1mW input signal, in the frequency range 300 to 6 000c/s, is fed to two emitter-followers VT_1 and VT_2. One of the outputs is amplified by a common emitter stage VT_3 and then split into four equal components using a purely resistive splitter to provide inputs for four of the balanced modulators in the Random Fading Unit.

The output from VT_1 is fed to a delay line for the simulation of frequency selective fading if required. The delay line consists of 70 basic sections of an inductor and capacitor switchable in blocks of seven sections enabling delay times from 0·2 to 2msec to be chosen. For convenience the switching is accomplished by ten toggle switches. The 1dB attenuator pad and equalization network ensure that the overall attenuation remains constant to within $\frac{1}{4}$dB for any desired delay time. The delayed output is amplified by VT_4 and split into four equal components to provide inputs for the other four balanced modulators in the Random Fading Unit. The delay line may be switched out of circuit by S_1 for flat fading conditions.

Random Fading Unit

This consists of eight balanced modulators each supplied with a separate carrier nominally at 100kc/s. The modulators are the single balanced switched type each employing four OA95 diodes. The 100kc/s carrier is generated by a crystal oscillator followed by a stage of amplification. This is shown for one of the eight modulators in Fig. 2. VT_5 and VT_6 form the oscillator with a 5° X-cut crystal connected between the emitters and operating in the series mode. Amplitude stability of the output is achieved by applying a.g.c., derived from the rectified output of VT_8, to the base of VT_5. The frequency of oscillation is adjustable over the small range $\pm 1{\cdot}5$c/s with a 3 to 30pF air dielectric trimming capacitor. The eight crystals and trimming capacitors are housed in the same enclosure so that the environment is the same for all eight circuits. The order of stability obtained is such that when the oscillators are set up with 0·05c/s separation, i.e. 10 fades/minute, the maximum drift of any one of the oscillators relative to the next is 0·02c/s over a period of one hour. The oscillators also maintain their relative frequency separations with a mains voltage change of ± 10 per cent.

Noise Generator

The noise is generated in an open-circuited emitter-follower stage VT_{14}, Fig. 3, and then amplified and selectively filtered to provide approximately 3mW into the 75Ω calibrated attenuator. The r.m.s. noise voltage measured at the collector of VT_{18} is approximately 1·2V. It is shown in the Appendix that the probability of the noise voltage exceeding ± 6V is less than 1 in 10^6. A 24V supply is used for VT_{18} and the circuit designed to give an undistorted voltage swing of 12V peak-to-peak. This is essential to avoid clipping of noise peaks.

The noise and fading signal are transformer coupled to the 100·3 to 106kc/s u.s.b. crystal band-pass filter.

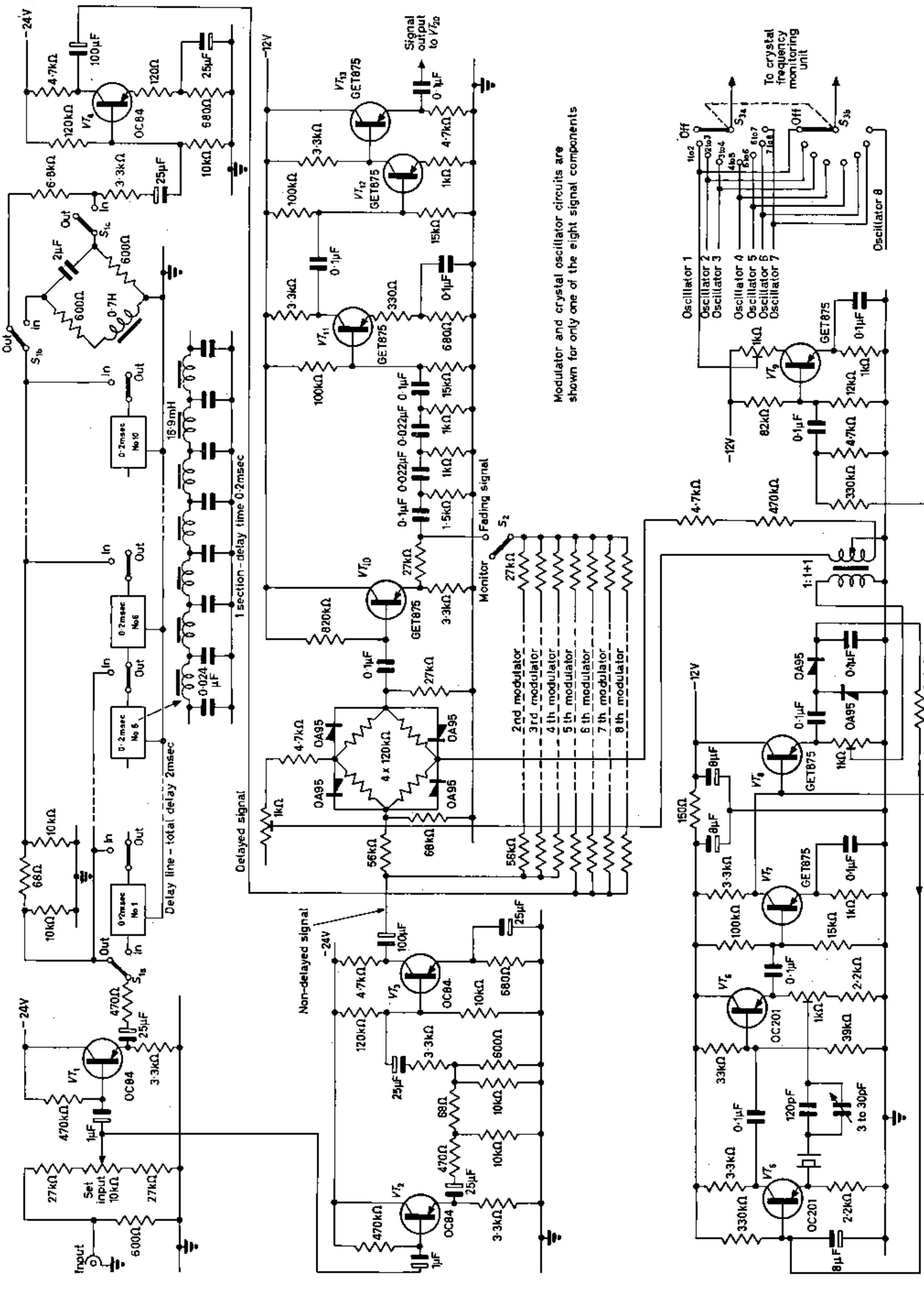


Fig. 2. Input stage and random fading unit

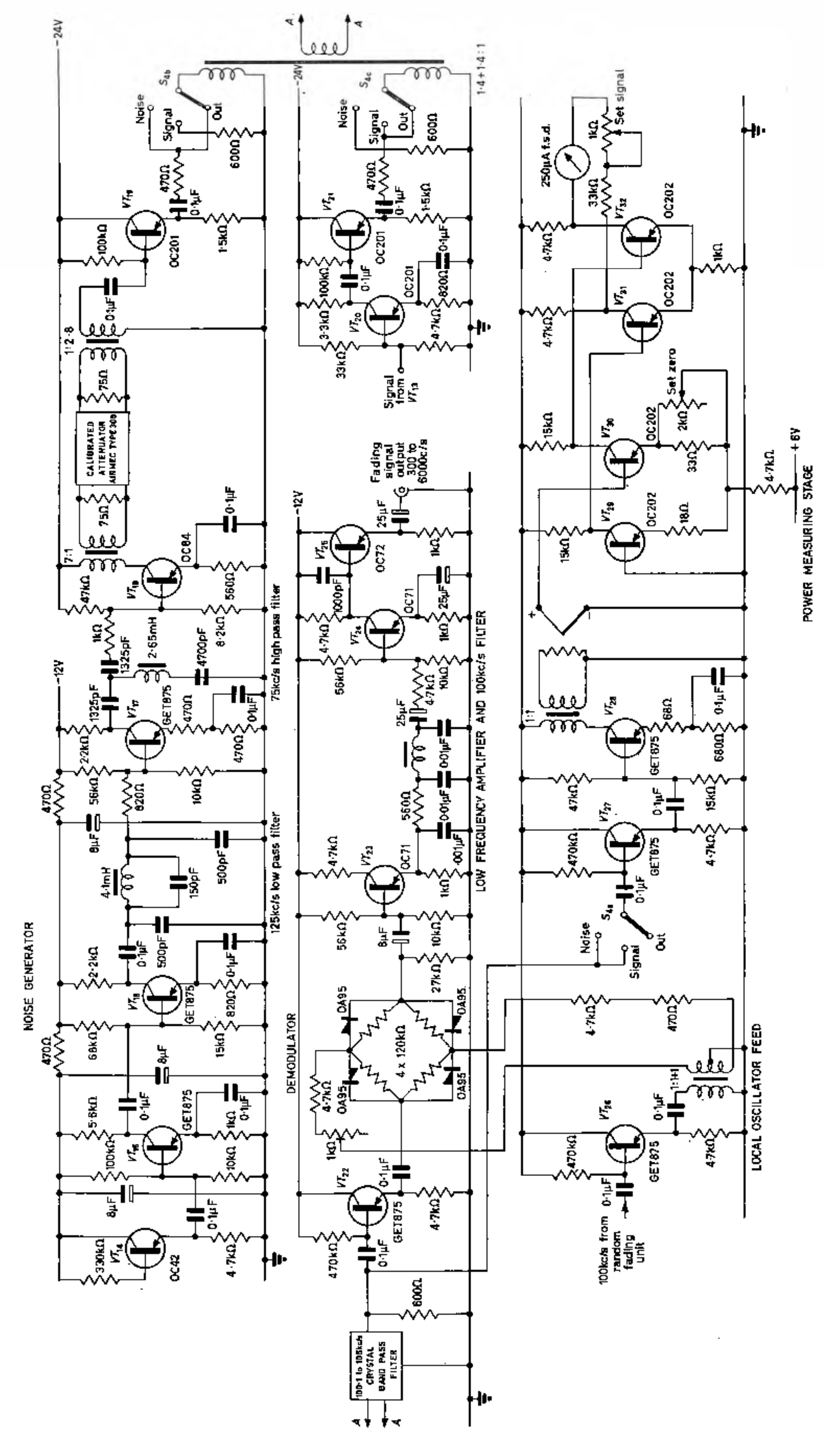


Fig. 3. Noise generator, power measuring and output stages

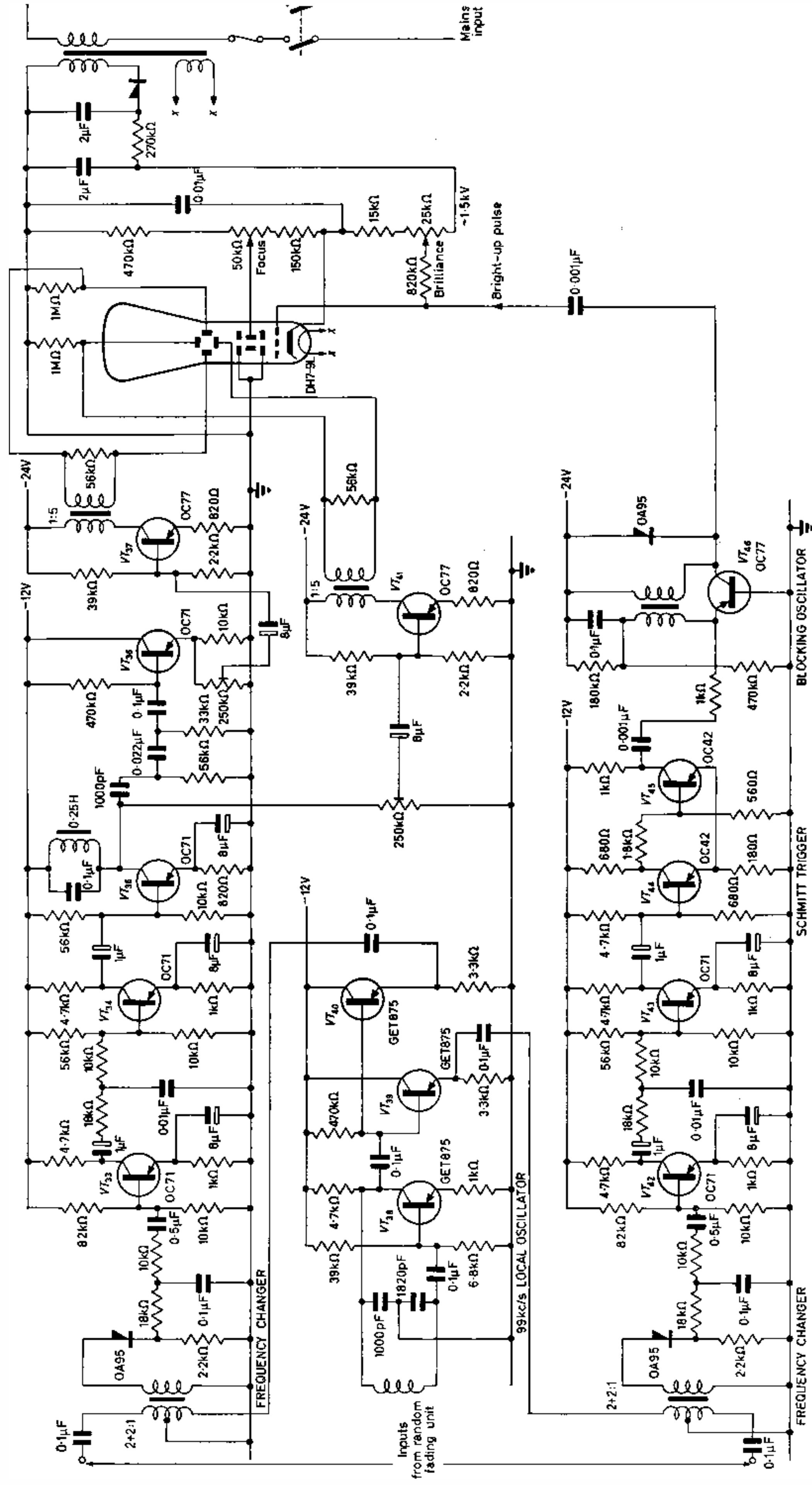


Fig. 4. Crystal frequency monitoring unit

Output Stage

The output from the u.s.b. filter is fed to an emitter-follower VT_{22} and then demodulated to restore the fading signal to the original frequency range 300 to 6 000c/s. The local oscillator for the demodulator is taken from one of the 100kc/s crystal oscillators in the Random Fading Unit. This is supplied to the emitter-follower driver stage VT_{26}. The demodulated output is amplified by $VT_{23\text{-}24}$. A single low-pass π section filter is included to remove any residual 100kc/s components. The emitter-follower VT_{25} provides an output of 8mW into a 600Ω load for the minimum stated input of 1mW.

Signal-to-Noise Ratio Measurement

The noise power is measured by switching out the signal with S_4 and amplifying the output of the u.s.b. filter with $VT_{27,28}$. The collector output is transformer coupled to the heater of an insulated thermocouple unit; a heater current of 1mA produces a 7mV open-circuit output voltage. The d.c. output from the thermocouple is amplified by two long-tailed pair stages $VT_{29,30}$ and $VT_{31,32}$. The differential output is displayed on a 250μA f.s.d. moving-coil meter.

The signal power is measured for only one of the eight components by switching S_2 in the 'monitor' position. A 9dB allowance is made when selecting a required signal-to-noise ratio for the fact that there are eight equal components forming the resultant fading signal.

The signal-to-noise ratio is set up by producing equal deflexions on the meter for both noise and one signal component. The calibrated attenuator in the noise circuit is then used to give the desired signal-to-noise ratio.

Crystal Frequency Monitoring Unit

The unit permits the setting up of the eight crystal oscillator frequencies for any desired fading rate. The oscillator frequencies are compared by switching pairs of adjacent oscillators to the monitoring unit. It must be remembered that only the frequency separation between the components is important in deciding the fading rate and that the absolute frequency need only be at a nominal 100kc/s.

The two inputs whose frequencies are to be compared are mixed with a local oscillator at 99kc/s and the difference frequencies of 1kc/s are extracted. One of the demodulated outputs is amplified by $VT_{33\text{-}35}$. A 90° phase shift circuit is included between VT_{35} and VT_{36}. In-phase and quadrature components are finally amplified by VT_{41} and VT_{37} respectively. The outputs are applied to the X and Y plates of the c.r.t. to form a circular trace.

The second 1kc/s demodulated output is amplified by $VT_{42,43}$ and applied to a Schmitt trigger stage. The sharp edge formed by the circuit is fed to a blocking oscillator VT_{46} to produce a positive-going bright-up pulse to the grid of the c.r.t. The resultant trace is a bright spot rotating at the difference frequency of the two input signals.

Conclusions

The fading of an h.f. signal is simulated for two path propagation with time differences of up to 2msec. The equipment could be modified for three path propagation and a longer time delay of 4msec if required.

It is hoped that the unit will form the basic tool for future investigations of improved techniques for h.f. telegraphy. Although considerable effort, by most research establishments, is being placed on satellite communication systems, the h.f. medium will always provide an essential link for some services particularly ship-ship and ship-shore communication.

Error detecting and counting equipment is now in hand to form a compact and transportable unit with the Fading Machine.

Acknowledgments

The Authors wish to thank Professor J. F. W. Bell and the Research Committee of the College for the facilities provided.

Appendix

The probability density function for a Gaussian distributed variable is given by:

$$p(x) = 1/\sqrt{(2\pi)}\exp\{-(x^2/2)\}$$

where x is the instantaneous value in units of the r.m.s. value of the function[5]. The probability distribution becomes:

$$P(x) = \int_{-x}^{x} p(x)\,dx$$

Tables of the error integral[6] show that the probability of the function exceeding $x = \pm 5$ is less than 1 in 10^6.

For an r.m.s. noise voltage of 1·2V the probability of the peaks exceeding $\pm$6V is therefore less than 1 in 10^6.

REFERENCES

1. Allnatt, J. W., *et al.* Frequency Diversity in Reception of Selectively Fading F.M. Signals. *Proc. Instn. Elect. Engrs.* *104*, 99 (1957).
2. Law, H. B., *et al.* An Improved Fading Machine. *ibid.* p 117.
3. McNicol, R. W. E. The Fading of Radio Waves of Medium and High Frequencies. *Proc. Instn. Elect. Engrs.* *96*, Pt. 3, 517 (1949).
4. Lord Rayleigh. Problem of Random Vibrations and Random Flights in One, Two or Three Dimensions. *Phil. Mag.* *37*, 321 (1919).
5. Baghdady, E. J. Lectures on Communication System Theory. p. 64 (McGraw-Hill, 1961).
6. Comrie, L. J. Six-Figure Mathematical Tables. Vol. II, p. 520 (Chambers, 1949).

A New Microwave Link Aerial

Scientists at the St. Mary Cray, Kent laboratories of Standard Telephones and Cables Ltd have produced a new low-cost aerial for the very-short-wave (microwave) radio links used for the transmission of television and telephone traffic.

The new £175 000 television/telephone link between London and France, now being built for the G.P.O., will be the first to use the new STC aerial.

The aerial uses an idea originally put forward in 1672 by Professor Cassegrain of Chartres University while working on the improvement of astronomers' telescopes. The principle involves double focusing to obtain sharper and clearer images.

As a result of this modern application of Professor Cassegrain's invention to microwaves, which behave like light beams and can be focused in the same way, the 29ft high 'hog-horn' aerials used on many radio towers can now be replaced with Cassegrain aerials of only 12½ft diameter without loss of performance.

The STC Cassegrain aerial is not only half the size but also one-quarter the cost of the equivalent hog-horn aerial. The reduction in weight also means that strengthening of tower structures is no longer required.

In the Cassegrain telescope a small convex mirror is placed near the focus of the main, large concave mirror. The received image is viewed through a hole in the large mirror. The incoming rays are therefore subjected to focusing and refocusing.

When applied to microwave aerials the radio energy is collected by a 12ft 6in diameter concave reflector and directed on to a 6in diameter convex reflector near the horn-shaped end of the circular waveguide that feeds the energy down the tower to the receiving equipment. The transmission process is exactly the reverse, the double focus action producing an extremely parallel and efficient microwave beam aimed exactly at the next radio tower, some 40 miles away.

The Field-Effect Transistor Used as a Low-Level Chopper

By K. Barton*

The field-effect transistor chopper is compared with the conventional silicon alloy transistor chopper. By using a field-effect transistor (f.e.t.) chopper, it is shown that a voltage drift of less than 0·1μV/°C, and a long-term drift of less 0·1μV/month can easily be achieved without expensive selection procedures, or the use of a temperature-controlled oven.

Included in the appendices are the calculations of forward gain of the f.e.t. chopper, for e.m.f. and current inputs, and the calculation of offset voltage due to offset current.

(Voir page 136 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 143)

THE performance of a low-level d.c. amplifier is usually specified in terms of noise referred to input. The high-frequency components of this noise can be filtered off, and do not become a serious problem unless the amplifier is required to have a wide band width. The low-frequency components (<1c/s) of the noise, present the greatest problem. These components are usually called offset, or drift, and can be split up into two parts, (a) voltage drift and (b) current drift. Drift is mainly caused by a change in temperature, or by ageing effects of the input circuit. For e.m.f. inputs, the error due to voltage drift is independent of source resistance, but the error due to current drift is directly proportional to it. The current- and voltage-drift figures are sometimes multiplied together and referred to as a power drift at the input. This is of very little practical value, however, and serves only as a figure of merit for the amplifier.

The Alloy Junction Transistor Chopper

The silicon alloy junction transistor makes an excellent switch, but has one major disadvantage. Since the transistor has to be driven 'on', a small voltage (e.g. 10 to 20mV) appears in series with it when it is closed, and the equivalent circuit is as shown in Fig. 1. If the transistor is used in the reverse connexion (i.e. collector used as the emitter), the voltage appearing in series with the closed switch is reduced to about 1 to 2mV. This voltage (usually called the 'pedestal voltage') can be backed off, but, unfortunately, it varies with temperature. Fig. 2 shows the typical variation of pedestal voltage with temperature at various base currents for a typical silicon alloy transistor. It can be seen from Fig. 2, that, for a particular transistor, a base current can be chosen of such a value as will ensure that the variation in pedestal voltage with temperature is at a minimum. If a variation of less than 2 or 3μV/°C is required, this technique cannot be applied to production amplifiers.

When drifts of the order of 0·2 or 0·3μV/°C are required, the transistor has to be maintained at a controlled temperature. This may be achieved by placing the transistor in an oven, in which the temperature is maintained to within 1°C or 2°C. A serious limitation is thus imposed upon the ambient-temperature rating of the amplifier. Also, the power required to heat the oven will not be less than 1 or 2W, which may be twice that required for the amplifier alone.

Long-term drift of the silicon alloy junction transistor chopper is unpredictable, and an expensive selection process is necessary if a long term drift of less than 5μV/week is required. Before selection, the transistors have to be aged for a few weeks at the temperature at which they are to be used. The ageing and selection process may take 6 or 7 weeks with a probable yield of about 25 per cent. Another disadvantage of the alloy junction transistor is the warm-up drift. The drift during the first two to three hours after switch-on can be as high as 50μV.

Fig. 1. Equivalent circuit of a silicon alloy chopper

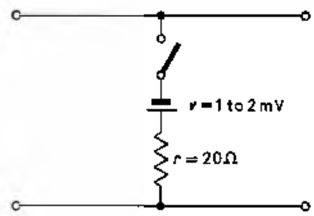
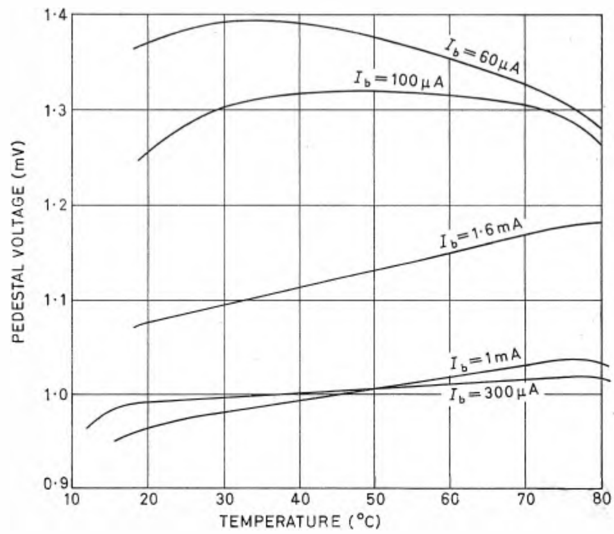


Fig. 2. Variation of pedestal voltage with temperature



Brief Description of the Field-Effect Transistor (f.e.t.)

The basic construction of the 'n' channel f.e.t. is as shown in Fig. 3. A bar of 'n' type material has two strips of 'p' type material alloyed into opposite surfaces. These form a pn junction. Ohmic contacts are made on to the two strips of 'p' type material which are then connected together. This is called the gate or control electrode. Ohmic contacts are then made on to each end of the 'n' type bar. One is called the source, and the other is called the drain electrode. It is, fundamentally immaterial which terminal is used for the drain or source electrode.

* *George Kent Ltd.*

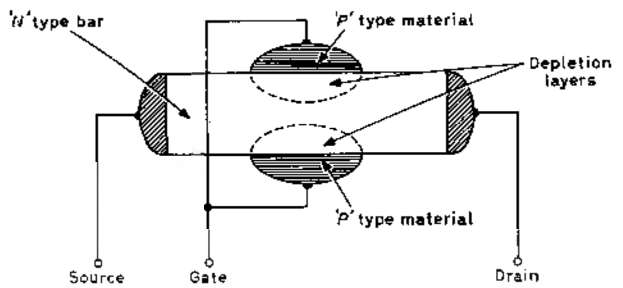


Fig. 3. Basic construction of f.e.t.

Operation

With no voltage applied to the gate (i.e. gate open-circuit) the slice of 'n' type material behaves as a pure resistance whose value depends upon the type of material and the construction of the f.e.t. Typical values of 'on' resistance of an f.e.t. suitable for chopping, are 300Ω to 2kΩ.

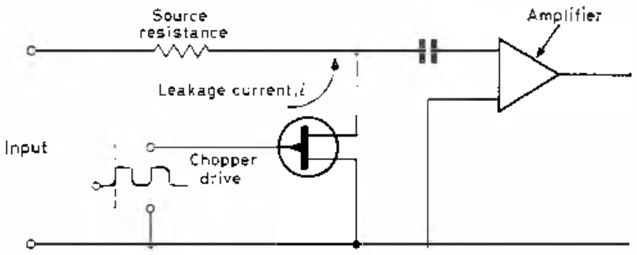
When a negative bias with respect to the source is applied to the gate, the junction is reverse biased and holes in the 'p' material and free electrons in the 'n' material migrate away from the junctions. Depletion layers are then projected into the 'n' type material (shown dotted in Fig. 3) causing an effective increase in the resistance of the bar. If the negative bias is increased so that the two depletion layers join, the channel is said to be pinched off, and the resistance between source and drain is practically infinite. So, by the application of the gate voltage, the resistance between source and drain can be switched between an open-circuit and, say, 300Ω.

The Use of the F.E.T. as a Chopper

With no voltage applied to the gate of an f.e.t., the resistance between source and drain is low (i.e. chopper closed). There is also no pedestal voltage developed in series with the closed resistance, and hence there can be no voltage offset. Such an arrangement is very suitable for chopping signals from low-resistance sources. To turn the f.e.t. off (i.e. chopper open) a gate voltage has to be applied, thus causing a leakage current to flow through the transistor in the 'off' state, as shown in Fig. 4. This current can be referred to the amplifier input as a current offset and an associated voltage offset (owing to the presence of the source resistance and the closed resistance of the chopper. See Appendix 2). The leakage current of a presently available f.e.t. such as the 2N2499 which is suitable for use as a chopper, is typically 5nA at 20°C and 22nA at 60°C. The voltage offset caused by this current, for a particular f.e.t. and a source resistance of 1kΩ, is about 20μV over the same temperature range.

The leakage current of a planar f.e.t., is almost perfectly logarithmic with temperature. If a planar diode is selected to have the same leakage current as the f.e.t., it can be used to prevent the leakage current from causing an offset at the amplifier input. Since the leakage current is

Fig. 4. The f.e.t. chopper in the 'off' state



present only when the chopper is open, it is desirable to find an arrangement which will ensure that the compensation current is there only when required. This can be achieved by driving the compensating diode from a square-wave source. Fig. 5 shows a 'p' channel f.e.t. chopper, with leakage-current compensation. With the arrangement shown in Fig. 5, a current drift of 0·5nA and a voltage drift of 0·5μV can be achieved over a temperature range of 20°C to 70°C.

To avoid the selection of a compensating diode, the amplitude of the square wave applied to the diode can be varied, or a diode used whose leakage is in excess of that of the f.e.t., and the current from it divided into two paths as shown in Fig. 6.

If the voltage drift caused by the compensation current flowing through the closed resistance of the chopper (150Ω

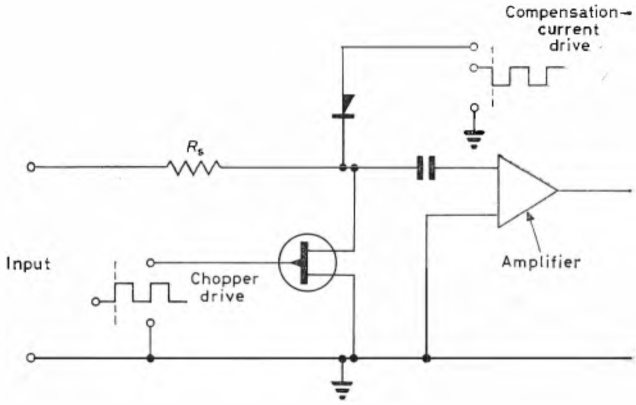


Fig. 5. The f.e.t. chopper with leakage-current compensation

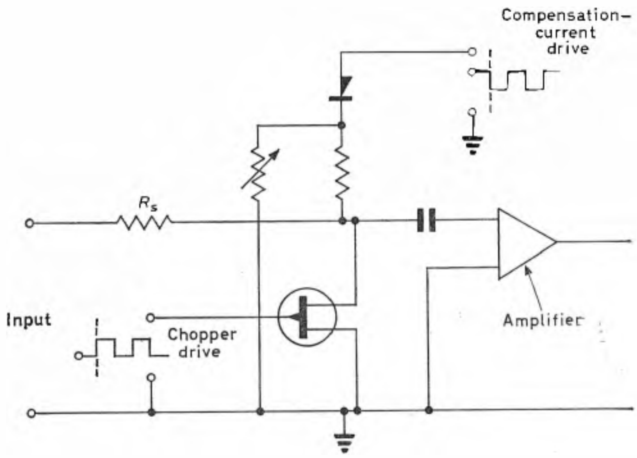


Fig. 6. Alternative method of leakage-current compensation

to 600Ω) can be tolerated, the diode can be returned to a d.c. supply, instead of being driven by a square-wave. The voltage drift to be expected in this case is between 5μV and 15μV over a 20°C to 60°C temperature range.

F.E.T.'s are now becoming available, although expensive, with leakage currents of less than 1nA at 65°C, and also very low junction capacitances. These transistors usually have relatively large values of 'on' resistance, which can be a disadvantage. Using these transistors, choppers can be made without the necessity for leakage-current compensations with possibly, an even better performance.

Chopper Spikes

Planar f.e.t.'s, which are available at present with low values of 'on' resistance (i.e. suitable for chopper use), have relatively large junction capacitances (e.g. 10pF to 20pF). The charge stored in these capacitances causes spikes which, when referred to the amplifier input, appear as a current offset. The magnitude of the current offset depends upon the amplitude of the chopper drive and the chopping frequency. At high chopping frequencies (3kc/s to 6kc/s) the spikes produced can be a serious problem. At a chopping frequency of, for example 200c/s the spikes cause an offset of about 2nA to 3nA. The change in this offset with temperature, and age etc. is, however, quite small. Provided the source resistance does not vary, this effect can be backed-off with the zero control. If the source resistance is allowed to vary, then account must be taken of the standing-current offset.

CHOPPER NOISE

The principal source of noise in a conventional transistor is due to carrier recombination in the base region. The f.e.t. is free from such noise because the current flow is determined by majority carriers, as in a metal conductor such as copper. So therefore, the f.e.t. has a noise figure at least an order better than a silicon alloy transistor. The random, low frequency (<2c/s) noise produced by the f.e.t. is typically $0{\cdot}2\mu$V to $0{\cdot}3\mu$V whereas the noise produced by a silicon alloy chopper is typically 2μV to 3μV.

WARM-UP DRIFT

The warm-up drift of a temperature-maintained silicon alloy chopper is usually quite large (e.g. 50μV in the first three hours). The f.e.t., however, does not drift in this way. The only drift that occurs from 'switch on' is due to thermal e.m.f.'s being generated in the input components as the amplifier begins to warm up. The magnitude of these effects is typically less than 3μV to 4μV and vanishes after 10 to 20min. Since the drift that does occur is normally within the amplifier specification, no warm-up time or ageing time is necessary.

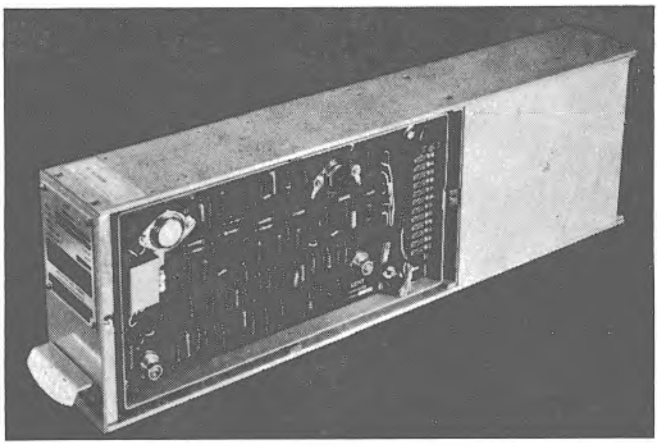


Fig. 8. Complete chopper amplifier using a silicon alloy chopper

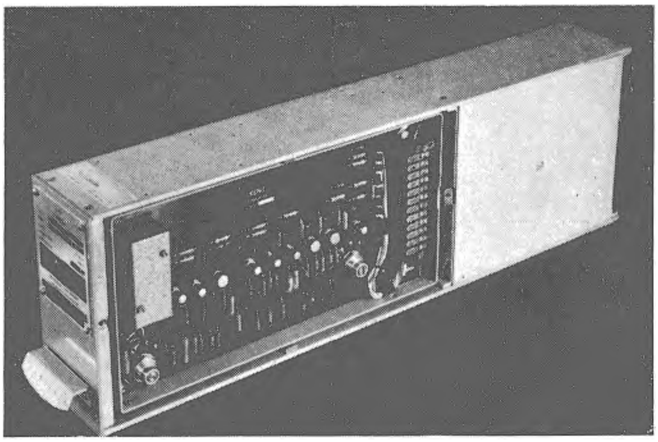


Fig. 9. Complete chopper amplifier using an f.e.t. chopper

LONG-TERM DRIFT

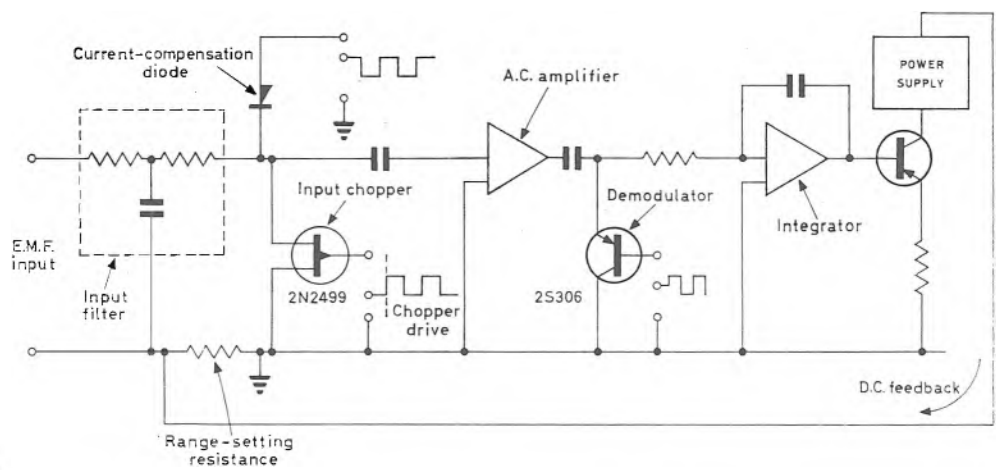
The pedestal voltage of a silicon alloy transistor, has to be accurately backed off with an equal and opposite voltage. The pedestal voltage appears to change owing to surface effects etc., at the transistor junctions, and varies in an unpredictable way. This effect and the stability of the 'backing off' voltage, limit the long-term drift performance of the silicon alloy chopper.

The long-term drift of the f.e.t. chopper is exceptionally good, since there is no pedestal voltage to be backed off, and the transistor is of planar construction. Drift rates of less than $0{\cdot}5\mu$V/month can easily be achieved. Drift rates of less than $0{\cdot}25\mu$V in nine months have been observed in some cases.

A block diagram of a complete f.e.t. chopper amplifier for e.m.f. inputs is shown in Fig. 7. The minimum span for the amplifier is normally 0μV to 100μV. But noise is sufficiently low to allow a span of 20μV if the change in ambient temperature is less than 5°C to 10°C.

Fig. 8 shows a chopper amplifier, using a silicon alloy chopper with oven, mounted in its case with range card and power supply. The top section of the board contains the oven-control amplifier, with the oven on the extreme left. Fig. 9 shows a redesigned version of the same amplifier using the f.e.t. chopper. It can be seen from Figs. 8 and 9, that there is a considerable saving in components effected by the use of the f.e.t. chopper.

Fig. 7. The f.e.t. chopper amplifier for e.m.f. inputs



APPENDIX

(1) GAIN OF CHOPPER

When calculating the loop gain of a chopper amplifier it is required to know the forward gain of the chopper. The amplifier briefly described above, fitted with a 50c/s filter on the input, was designed to work from a source resistance which could vary between 0 and 1kΩ. The gain of the chopper from input to output for current and e.m.f. inputs is given below.

For e.m.f. inputs, the gain is defined as: The peak-to-peak output current divided by the direct input voltage. For current inputs, the gain is defined as: The peak-to-peak output current divided by the direct input current.

E.M.F. Inputs

Fig. 10 shows the input circuit with RC filter and using an f.e.t. chopper where r is the 'on' resistance of the chopper, R_3 is the input resistance of the a.c. amplifier, and R_s is the source resistance.

For e.m.f. inputs, the gain of the above circuit is as follows:

$$\text{Gain} = \frac{G.Z.}{(R_s + R_1 + Z)} \text{ amperes/volt}$$

$$\text{Where } G = \frac{1}{(R_3+r\,.(R_3/R_2)+(R_2/2)+r)} \text{ amperes/volt}$$

$$\text{And Z} = \left(2r + R_2 + \frac{R_2R_3}{R_2 + R_3}\right) \text{Ohms}$$

Example

With an amplifier input resistance $R_3 = 600\Omega$ and a chopper "on" resistance $r = 600\Omega$.

Then $G = 487\mu A/V$

and $Z = 2570\Omega$

e.g. With $R_s = 0$

Gain $= 350\mu A/V$

e.g. With $R_s = 1k\Omega$

Gain $= 244\mu A/V$.

When calculating the effect of series-mode pick-up etc., it is useful to know the gain from the capacitor C onwards. This is equal to G i.e.

$$\frac{1}{(R_3+(rR_3/R_2) + (R_2/2)+r)} \text{ amperes/volt}$$

The input resistance in the steady state, i.e. for direct input voltages is:

Input resistance $= R_1 + Z$

$$\therefore \text{ The input resistance} = R_1 + 2r + R_2 + \frac{R_2\,R_3}{R_2 + R_3} \text{Ohms}$$

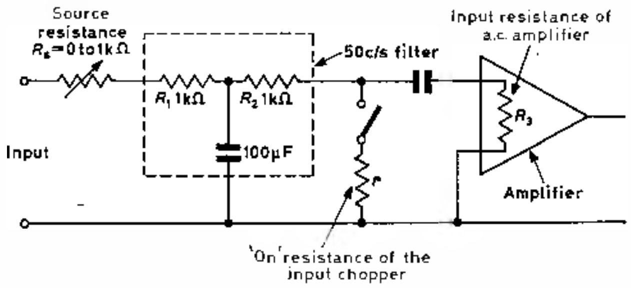
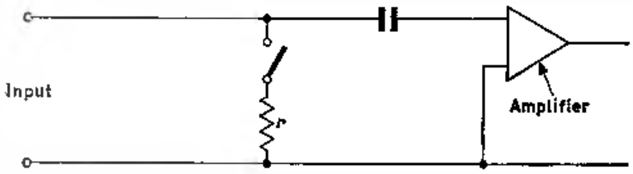


Fig. 10. Input circuit of f.e.t. chopper for e.m.f. inputs

Fig. 11. Input circuit of f.e.t. chopper for current inputs



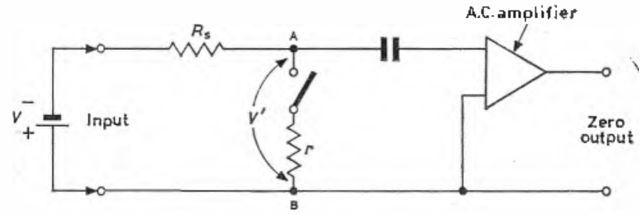


Fig. 12. Equivalent circuit of the f.e.t. chopper in the 'on' state

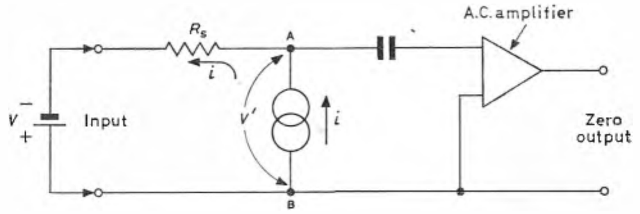


Fig. 13. Equivalent circuit of the f.e.t. chopper in the 'off' state

Current Inputs

Fig. 11 shows the effective input circuit for current inputs. The gain of this circuit for a 1:1 mark-space ratio is practically independent of the 'on' resistance of the chopper when the 'off' resistance is much greater than the amplifier input resistance, and is as follows:

Gain for current inputs $= 2$

(2) VOLTAGE OFFSET OF THE F.E.T. DUE TO CURRENT LEAKAGE

When the f.e.t. chopper is in the open state, the gate voltage causes a leakage current to flow between gate and drain. This current gives rise to an effective voltage offset due to the presence of the source resistance and the closed resistance of the chopper. An equation is derived below giving the magnitude of this off-set voltage in terms of the circuit parameters. Fig 12 shows the equivalent circuit of the chopper in the 'on' state. With no input applied, V will be zero. Fig. 13 shows the equivalent circuit of the chopper in the 'off' state. With no input applied the effect of i flowing through the source resistance R_s will cause an offset at the amplifier input.

If there is to be zero output from the a.c. amplifier (i.e. no a.c. signal fed into the amplifier) then the voltage between points A-B must be the same in the 'on' and 'off' states of the chopper. To achieve this, an input V is applied as shown. This input is the effective voltage off-set of the chopper.

Let R_s = Total source resistance

r = 'On' resistance of the chopper

i = Gate-drain leakage current of the f.e.t.

V = The effective voltage off-set caused by the leakage current

If the voltage between A-B is to be the same during both half cycles then $\frac{V.r}{(r + R_s)} = V - i.R_s$.

$\therefore$ The voltage off-set $V = i\,(R_s + r)$ volts.

REFERENCES

1. HUTCHEON, I. C., SUMMERS, D. A Low-drift Transistor chopper-type d.c. Amplifier with High Gain and Large Dynamic Range. *Proc. Instn. Elect. Engns.* *107 B*, 451 (1960).
2. CHAPLIN, G. B. B., OWENS, A. R. Some Transistor input Stages for High Gain D.C. Amplifiers. *Proc. Instn. Elect. Engns.* *105 B*, 249 (1958).
3. CHAPLIN, G. B. B., OWENS, A. R. A Transistor High Gain chopper-type D.C. Amplifier. *Proc. Instn. Elect. Engns.* *105 B*, 258 (1958).
4. HUTCHEON, I. C. The Contact Modulator (part 3), Modulation Methods. (Airpax Electronics Incorporated, Florida U.S.A. and G. Kent Limited 1959).
5. G. KENT LIMITED, Patent Application No. 39474/63.

Fixed Resistors for Use at High Voltages

By P. L. Kirby*, D.Sc., F.Inst.P.

Means by which the geometry of a film resistor can be controlled to minimize voltage gradients are considered. The problem of non-linearity and the limitations of a 'voltage coefficient' are discussed. For high voltage applications it is convenient to combine voltage and temperature coefficient effects. An improved form of high voltage resistor based on a thick film of 'diluted' tin oxide is described. The noise level is remarkably low and the resistors are stable to about ±1 per cent under normal conditions and can tolerate voltage pulses of up to 50kV peak. Spark gaps can be provided as a protection against higher voltage surges. The behaviour of these resistors in the bleeder chain of a high voltage electrostatic generator is described.

(Voir page 136 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 143)

THE maximum severity of the electrical loading condition which can be applied to fixed resistors used in electronic circuits is usually limited by the maximum power which can be dissipated by the component. In those cases where a high value resistor is used, a maximum working voltage comes into operation and the resistance value where both maximum power and maximum voltage can be applied is usually termed the 'critical value' ($R_c = V_{max}^2/W_{max}$).

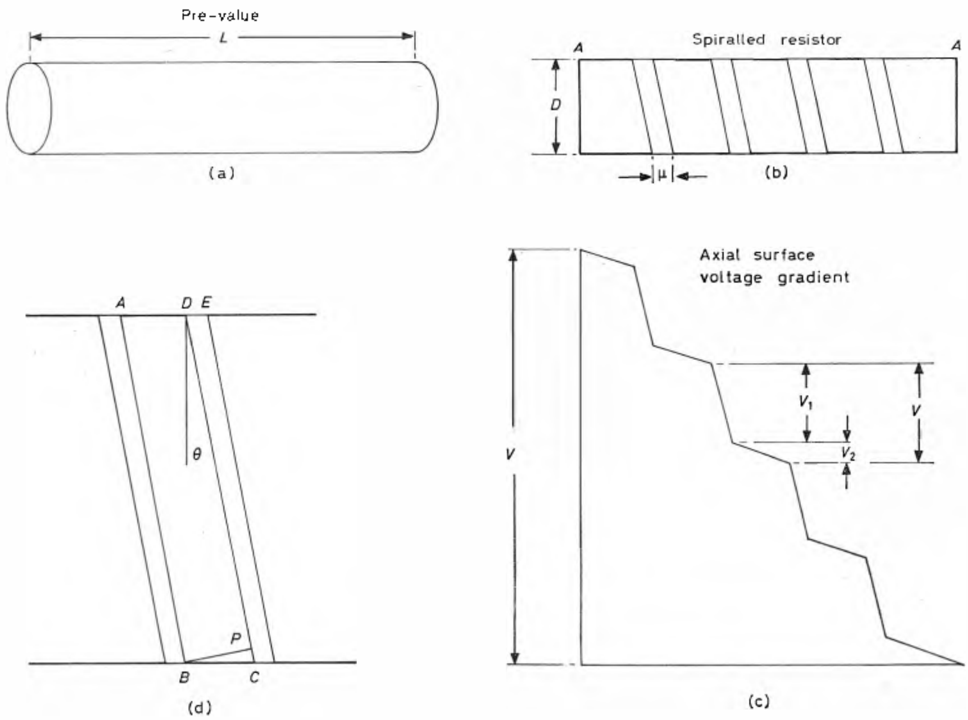
For the majority of small fixed resistors the maximum voltage is found to be of the order of 1kV per inch of resistor length. This approximate criterion applies to film resistors, where the conducting track lies on the surface of an insulating core and a helical groove is cut through the film to extend the length of the resistive path, and also to solid composition resistors. The use of similar maximum voltages might seem surprising as, even if breakdown did occur, quite different mechanisms would be involved in the two cases. The solid resistor would undergo internal intergranular breakdown and the film resistor would exhibit surface tracking. In fact, however, maximum voltages are much lower than the breakdown level and are usually applied to reduce the power dissipated in high value resistors of all types, to ensure an operating stability similar to that achieved with lower value resistors at full dissipation.

At one time the general rugged construction of solid composition resistors compared with the frail and one-time low reliability of film resistors led to a general preference for the former if the application involved higher than normal voltages. Some types of solid composition resistors will fail under high voltage conditions (provided the maximum current is limited) by degrading initially to a very low value. For some non-critical applications this mode of failure on the part of a small number of elements in a long resistance chain is not catastrophic. On the other hand any appreciable number of such failures will increase the current drain and may lead to a runaway condition with catastrophic results.

Nowadays the situation is very different and this article describes one form of high voltage film resistor in which the film is remarkably rugged and reliable. In addition the long helical track of a film resistor assists in the achievement of the high resistance values which are so frequently required in high voltage applications. This can only be achieved with a solid resistor by considerable 'dilution' of the conducting system with insulating materials, accompanied by a deterioration in other electrical properties such as noise or temperature and voltage coefficients.

It is appropriate therefore to consider whether there are any features of the design of a film resistor which require special attention if the resistor is to be used for high voltage applications. When a film resistor is designed particularly for use at high voltages it will usually be provided with a protection system which minimizes the likelihood of tracking over the surface of the resistor. In addition, particular aspects of the track geometry can be adjusted to provide optimum conditions for high voltage operation.

Fig. 1. The voltage gradients developed across the surface of a spiralled film resistor



* *Welwyn Electric Ltd.*

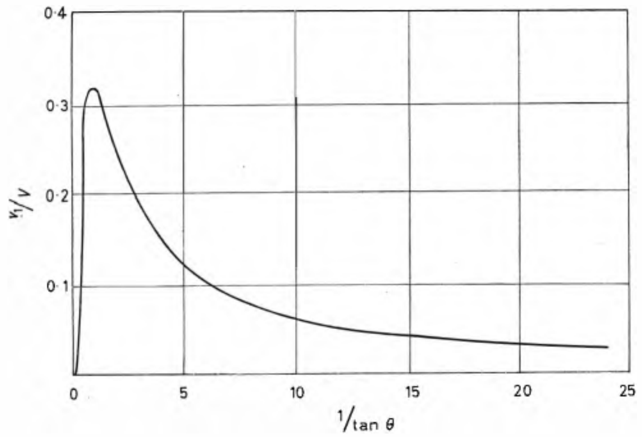


Fig. 2. The ratio of the voltage across the helical cut to the total voltage applied to the resistor as the number of turns is varied

The Geometry of a High Voltage Film Resistor

In the simple analysis of the 'gain' (G) in resistance which results from the application of a helical cut of pitch T turns/in and width μ inches along the cylindrical surface of a film resistor (diameter D) the terminal areas are ignored and the ratio of the final 'spiralled value' to the original 'pre-value' is given approximately by:

$$G = \frac{(T\pi D)^2}{1 - \mu T}$$

This equation serves adequately to express the maximum gain achievable for all the conditions normally encountered in the 'spiralling' operation. The helix angle (θ) is given by:

$$\tan\theta = 1/(T\pi D)$$

When high voltages are to be applied across the ends of the resistor it is important to consider two particular distributions of potential. On the one hand there is the uniform gradient along the helical track in the direction of current flow. This gradient is inversely proportional to the length of the track (l) which is related to the resistor length (L) by:

$$l = L\,.\,T\pi D$$

Secondly there is the distribution of potential lying axially along the surface of the resistive fim, intersected by successive turns of the helical cut and having therefore a non-uniform distribution. This distribution includes the highest local gradient occurring at any point on the resistor and is therefore of considerable importance. The situation is illustrated in Fig. 1.

The axial gradient along AA' with a constant voltage V across the resistor is illustrated in Fig. 1(c). The mean gradient is V/L and is equal to the surface axial gradient had the same voltage been applied to the pre-value. The latter represents the case where the actual in-line gradient along the track is a maximum.

If the voltage across any cut is denoted by v_1, and across the track by v_2, then the voltage which occurs across any complete turn of the spiralled resistor, $v = v_1 + v_2$, is inversely proportional to the number of turns, LT.

Thus, $v = V/LT$.

The partition of v between the cut (v_1) and the track (v_2) is deduced by geometrical considerations from Fig. 1(d), which represents a planar view of a single turn unwound from the resistor. The points B and E are thus identical and are also at the same potential as P. Thus:

$v/AB = v_1/DP = v_2/PC$ and it can be shown that

$$v_1/v = 1 - (1 - \mu T)\sin^2\theta$$

The ratio of the voltage across any cut to the total voltage applied to the resistor is therefore:

$$v_1/V = 1/L\,\frac{(\pi^2 D^2 T + \mu)}{(1+\pi^2 D^2 T^2)}$$

Fig. 2 shows how this expression varies with $1/\tan\theta$ (and hence with T, for constant resistor dimensions and cut width). The expression is zero for $T = 0$ (representing the case of a pre-value or a resistance track formed by a straight axial stripe where indeed there is no voltage drop *across* the track or the cut).

A maximum value of v_1/V occurs and if the width of the cut is assumed to be small compared with the track width (and it certainly is for low values of T) the maximum occurs approximately at $T = \pi D$. This corresponds to the rather unlikely practical situation of the helix angle, θ, being 45°. At higher values of T the term v_1/V does not vary appreciably. A maximum value of $\tan\theta$ can be stipulated so that v_1/V can be kept well below its maximum value.

It is appropriate to note that an additional advantage is gained by the use of a wide helical cut, especially at low values of T where there is less danger of reducing the residual track width to dangerously low levels.

Criteria such as a minimum permissible number of turns are applied in the design of high voltage film resistors. This particular safeguard also ensures that the uniform in-line gradient around the track will have been reduced and the element is working under voltage stresses very much lower than would apply with the solid form of construction.

Non-linearity and its Problems

In the use of fixed resistors for precision applications at high voltages, the non-linearity in the current-voltage characteristic can present serious problems. The reduction in voltage gradient inherent in the spiralled film resistor results in a considerable diminution of the effect but it is appropriate to consider the general aspects of the problem. The parameter 'resistance' which is derived from the current-voltage relationship will not be constant if the relationship is non-linear. It is not to be expected, however, that the deviation from linearity will be so simple that the resistance, in its turn, will vary linearly with the voltage and permit the use of a first order 'voltage' coefficient'. This represents a weakness in the concept of a voltage coefficient which in practice, is very dependent on the voltage range over which it is measured. A typical high voltage resistor may have a voltage coefficient of $-0{\cdot}05$ per cent/V over the range of 10 to 100V but show a change of only -1 per cent between 100V and 10kV (i.e. a coefficient of $-0{\cdot}0001$ per cent/V). The disparity between these two values of the voltage coefficient of one resistor illustrates how misleading this parameter can be.

For some applications a potential dividing network is required to provide a fixed ratio over a range of applied voltages. If the two resistive elements of the system have the same voltage coefficient this requirement will certainly not be met. The coefficient is defined as the percentage change *per volt* and as the voltage across the two elements is likely to be quite different then the 'percentage changes' will differ and constant ratio of voltages will not be achieved.

Finally, it is extremely difficult both as a test measurement or under practical operating conditions to distinguish variations in resistivity which result from a true non-linearity and those which occur due to the heating of the resistor and its temperature coefficient. Attempts to measure 'instantaneous' resistance changes when voltages are applied, or to use pulse techniques have not met with complete success. The practical difficulty of distinguishing between the two effects becomes even more pronounced when local heating is considered. For example, heating at intergranular contacts or even intercrystalline boundaries might occur and would be extremely difficult to dissociate from a 'true' voltage non-linearity. The use of a third harmonic index to describe the non-linearity occurring at low voltages is advantageous but a problem at high voltages still remains. Probably the most satisfactory practical approach is to combine the changes due to voltage coefficient and temperature coefficient and characterize the resistor in given ambient conditions when equilibrium under the electrical load concerned has been achieved.

Close Tolerance High Voltage Resistors

Occasionally there is a requirement for a very high value resistance to be used in conjunction with a voltmeter to monitor the voltage developed at some point of a high potential system. Similarly, resistors of accurately known value are required in the stabilizing circuits of high voltage apparatus. In these cases where the voltage is not likely to vary over a wide range, the problems imposed by the voltage coefficient of the resistor can be surmounted and it is possible to assemble a chain of fixed resistors in series such that the total resistance value under given operating conditions can be known to a high order of accuracy. It is merely necessary that the resistance of each individual component in the chain be determined under test voltages equivalent to that which will exist across each component in the final application. It is not possible to monitor the resistor at very high test voltages during the course of manufacture, and a convenient method, is to manufacture most of the resistors to a nominal value and then to calculate the value of the additional resistors needed to make up the rest of the chain after the first batch has been measured under operating conditions.

An Improved Type of High Voltage Resistor

There has always been some difficulty in selecting the best type of film resistor for use in the bleeder chains required for voltage equalization in high voltage electrostatic generators. The increase in maximum operating voltage of some of the larger machines has highlighted the deficiencies of existing resistors. It has been common practice to assemble fixed resistors of conventional size into convenient sticks between adjacent voltage equalizing planes and a very large number of individual elements are included in the whole chain. It is generally found that any type of solid composition resistor exhibits a large negative voltage coefficient and drops considerably from its nominal value as soon as the working voltage is applied. Progressive and large irregular subsequent changes in value are frequently observed.

A partial improvement may occur when resistive materials involving mixtures not dissimilar to those used for solid composition resistors are applied as a film on the surface of an insulating substrate. In the case of such film composition resistors the voltage gradient is reduced by the conventional use of a helical cut through the film to give a greatly increased length of track. Film resistors involving more conventional materials (metal alloys or oxides) may at first appear to provide a more stable performance but under the severe conditions obtaining in high voltage electrostatic generators, their reliability may not be adequate.

During an investigation[1] of the behaviour of high value fixed resistors in the bleeder chain of a 6MV tandem Van der Graaf generator it was found that the prime cause of resistor failure was the occasional occurrence of a high voltage transient pulse. Initiated by a flashover from the high potential electrode to the outer casing of the machine, a high voltage step appeared at one end of the resistor chain and this was transmitted at great speed down the whole stack. Because of the high velocity of propagation a very high potential gradient could occur for a short space of time across any given length of the resistor track. Due to the attenuation of the voltage step during its passage down the chain the resistors at the high voltage end of the uppermost stick suffered the most damage.

It was necessary to construct laboratory apparatus to provide a short high voltage pulse and for most tests a pulse of 50kV peak was applied to single fixed resistors varying up to about 2in in length. It was realized that the conditions within the electrostatic generator when flashover occurs are probably more severe. On the test apparatus it was established that there were several modes of failure of film resistors under high voltage pulse conditions. Even in those cases where a complete cap to cap flashover is avoided the resistive track could suffer damage. It was found that with film resistors incorporating a helical cut a more localized discharge or scintillation could produce changes in resistance. The magnitude of the peak voltage which produced this effect could be very greatly increased by the provision of an insulating coating of high dielectric strength around the film. When a 'pre-value' or non-spiralled composition film resistor was tested the high voltage gradient appeared to be responsible for changes in resistance which may have arisen from some internal form of breakdown.

Attempts were made to utilize high-quality film resistors whose basic stability under normal conditions had been proven and to ensure adequate performance under high voltage pulse conditions by the application of an insulating coating of high dielectric strength. Evidence was obtained to suggest that this approach was successful in eliminating forms of surface discharge which would certainly have damaged the resistor. Unfortunately, in order to obtain the high resistance values required (about 40MΩ per linear inch of resistor stick) rather thin films of the conventional materials (deposited carbon, oxide films, or metal films) were required. The tests showed that such resistors did not possess adequate reliability and that progressive increases in value with subsequent open-circuiting occurred too frequently.

Fortunately at this time the potentialities of a new type of resistive material were being assessed. The remarkable reliability of the tin oxide film resistor at low values, where the thick inorganic film with its physical hardness and chemical inertness assisted in its withstanding unusually severe conditions, led to thoughts of possible means of its further exploitation. Some attempts had been made to compound doped tin-oxide (i.e., in a conducting form) with other inorganic materials to form a new solid composition resistor. There is no report to date of this line of development having led to a satisfactory product. On the other hand, conducting tin oxide in association with inert diluents and inorganic binders applied to the surface of an insulating ceramic core had been shown to have useful properties as a fixed resistor[2]. The film, so applied, was particularly thick and very resistant to both physical and chemical damage. The surface resistivity (in ohms per square) was at least two decades higher than the highest permissible value of 'pure' tin oxide film, and the electrical stability of the new resistive material appeared to be extremely good.

As the 'natural' end point of this new development was to produce relatively high value films (having 'spiralled' values of from 1MΩ upwards) it seemed reasonable to aim the component at high value and high voltage applications.

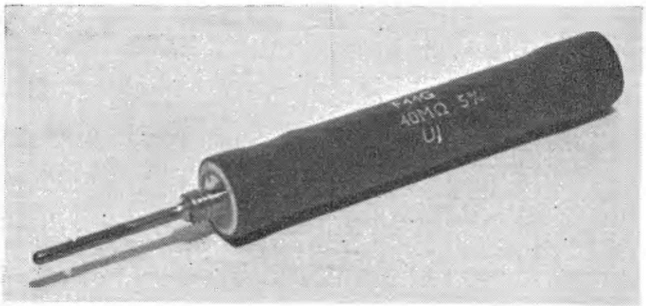
The question then arose as to whether, if this thick, high resistivity, inorganic film composition resistor was appropriately covered with an insulating coating of high dielectric strength, the component would provide an improved performance under the high voltage pulse conditions concerned in the present investigation. The short answer was that it did, and there emerged the high voltage Metox resistor in which the standard form consists of a rugged high resistivity inorganic film based on tin oxide and bonded to a high quality ceramic insulator at temperatures in excess of 1 100°C. The resulting film is extremely hard (it cannot be scratched with a steel blade) and is finally protected by the application of silicone rubber. The latter is renowned for its excellent dielectric strength and resistance to tracking and it is firmly bonded to the resistive film to minimize the danger of interfacial tracking. The latter is a common final mode of failure of the present product, but the ability of a fixed resistor of total length less than 2in to withstand repeated voltage pulses considerably in excess of 50kV speaks well for the general electrical worthiness of the assembled unit.

Additional Improvements

The high voltage Metox resistor (Fig. 3) represents a significant improvement in the field of fixed resistors for high voltage operation. It is still found, however, that under the extreme conditions which occur occasionally in high voltage electrostatic generators even these improved resistors can be affected detrimentally. As noted above, one ultimate failure mode initiated by the severe high voltage pulse (probably in excess of 100kV over the length of one resistor unit) is an interfacial tracking between the silicone rubber and the resistive track. A dark tree-like pattern develops on the surface of the resistor which is found to have increased in value, although the change in electrical properties is not confined to the particular area where the visible effect has occurred.

It is very unlikely that any form of fixed resistor made from any known materials could withstand the ultimate conditions of severity which can occur in the high voltage machines. At the present time it is a question of using the component which gives the longest life expectation and whose general electrical performance is in accord with the normal requirements of the equipment.

Fig. 3. High voltage Metox resistor showing the protruding electrode of an internal spark gap. The other electrode lies inside the resistor body and the 'gap' is formed when two resistors are screwed together



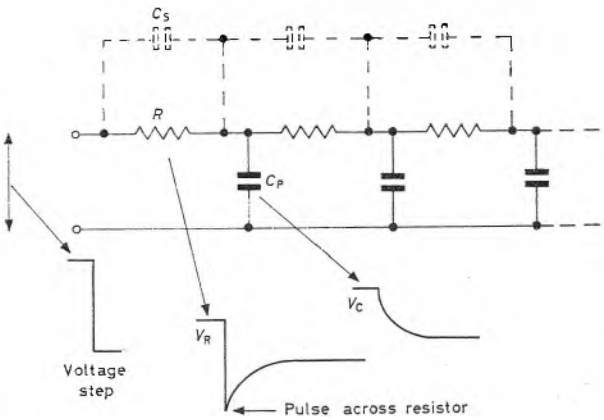


Fig. 4. The equivalent circuit of a bleeder chain in a high voltage electrostatic generator. The dotted lines indicate the additional capacitances which are provided by the external screening cans

The type of electrostatic generator concerned in the early investigation included spark gaps set between adjacent equipotential plates and these gaps would break down right down the stack when a flash-over took place. One must presume that the velocity of propagation of the voltage pulse down the resistor stick is so rapid that severe gradients are set up, at least across the first resistor element of the stick before the protective spark gap breaks down. This has led to the protection of each resistor by the application of internal electrodes to provide individual spark gaps for each resistor. The spacing is such that, in whatever gas atmosphere the unit operates, the internal spark gap will break down at a voltage in excess of the normal working level but below that at which damage would occur to the resistor. Tests have been carried out to ensure that corona discharge between the electrodes at the operating voltage is in no way significant compared with the steady bleed current through the resistor.

One likely reason for the appearance of a voltage 'pulse' across the resistor is the appreciable capacitance, C_p, to earth (or to the nearest equalizing plane) of any point along a resistor stick (Fig. 4). Thus when a sudden 'step' in the voltage at one end of the stick occurs the stray capacitances will not permit rapid equalization and a short 'pulse' will appear across the resistor. Some attempt might be made to balance this condition by an artificial increase in the shunting self-capacitance of the resistor elements. If the shunt capacitance can be increased so that C_sR is long compared with the rise time of the voltage step then the pulse in the V_R waveform will be eliminated.

So far as is known there has been no serious attempt to provide high voltage capacitors along the resistor chain to provide the required balance. It has, however, been possible to add an external screening can (Fig.5) to the resistors in an attempt to increase the self-capacitance. The opportunity is also taken to provide further external spark-gaps between one screening can and the next to provide additional protection to the individual resistors.

The combination of internal and external spark gaps and the provision of additional self-capacitance has produced a form of resistor far more capable of withstanding the severe conditions of flash-over in a high voltage generator than any previous component.

The external screening cans provide additional useful features. When they are incorporated in a high voltage resistor chain they ensure that each individual resistor is operating in an electrostatic field little greater than that defined by the voltage across one component. Any danger of corona discharge associated with the units at the high voltage end of the chain is minimized by the smooth large-radius outer contours of the screening can. It is also well known that whereas under stable d.c. operating conditions it is relatively easy to maintain uniform grading along a chain by the use of linear resistive elements, it is less easy to maintain uniformity under transient conditions. In this case it is the capacitances of the system which determine the effectiveness with which uniform grading is maintained. The external screening cans are therefore useful in adding uniformly to the self-capacitances of a resistance chain and thereby increasing the likelihood of maintaining voltage grading during a transient voltage surge.

High Voltage Metox Resistors

The present range of values of these resistors is from 2MΩ up to 10kMΩ at an accuracy of ±2 per cent. Lower value resistors (which may be required to permit low voltage tapping from a high voltage chain) utilizing the same basic materials and therefore having a similar temperature coefficient etc., to those in the normal range, are available at wider tolerances (±10 per cent) by the use of unspiralled pre-values.

The normal size of resistor (F44, 2in × $^{5}/_{16}$in diameter) is capable of 1W dissipation at values below 200MΩ. At higher values a maximum continuous d.c. voltage of 14kV can be applied.

The temperature coefficient has a mean value of 0·15 per cent/°C over the whole range of values. Shelf stability and changes on voltage load are less than 1 per cent and although high voltage resistors are not normally used in humid conditions these resistors are found to return closely to their original value if they are allowed to recover and dry off after exposure to the usual climatic test conditions.

The insulation between the resistive element and a neighbouring conductor is limited by breakdown from the lead across the end face of the resistor and occurs in the standard form at about 3kV. These components would normally be mounted in a stand-off manner so that full use can be made of their basic insulation. Breakdown from an end terminal to a test guard ring wrapped around the centre of a resistor occurs at voltages approaching 30kV.

The insulating protection applied to the resistor mounted in free air is adequate to withstand all the electrical conditions which the basic resistive element can withstand and thus there is no essential need for further insulation by immersion in a high pressure gas, or in oil or by encapsulating in resin. On the other hand, high voltage resistors are sometimes required to operate in environments dictated by the requirements of associated apparatus. High voltage Metox resistors operate satisfactorily in vacuum or high pressure insulating gases, it being necessary only to ensure that a form of construction is utilized which involves open access to the interior of the ceramic tube to which the film is bonded. For use in organic fluids the resistor can be supplied without the silicone rubber protection if there is any doubt about the long-term effects of the fluid on the rubber. As the resistive element is of such a rugged nature, there is no danger of damage to the unprotected unit during normal handling.

Fig. 5. A chain of high voltage resistors incorporating both internal spark gaps and external screening cans. The latter are arranged to provide additional external protective spark gaps

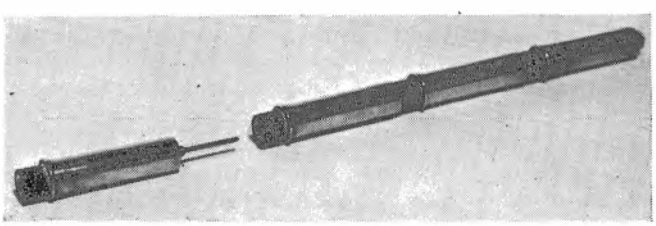


TABLE 1

Average Noise Level of High Value Resistors of Similar Overall Dimensions

VALUE (MΩ)	HIGH VOLTAGE METOX	DEPOSITED CARBON	CARBON/ RESIN FILM	SOLID CARBON COMPOSITION
10	−8	−4	−1	+16
20	−2	+1	+1	Not available
50	−2	+13	+5	,,
100	+4	+16	+15	,,
200	+8	—	+16	,,
500	+16	Not available	—	,,

The units are decibels referred to 1dB = 1μV/V in a decade

Finished resistors have been successfully potted or encapsulated in epoxy resins and this form of final assembly does, of course, offer useful advantages in the production of compact multiple assemblies for voltage dividing networks etc.

Current Noise in High Voltage Resistors

A high value fixed resistor is sometimes required as a load resistance for a high impedance thermionic device such as an ionization gauge or a photocell. In such applications the resistor must exhibit good stability so that the calibration of the system can be maintained and also, due to the likely need for further amplification of the output from the device, the load resistor will be required to have a particularly low noise level.

A wire-wound resistor does not produce any noise in excess of the random thermal fluctuations of the conductor and would, from this viewpoint, be an ideal choice. Unfortunately, high value wire-wound resistors are bulky, they have large values of inductance and capacitance and they are very expensive. Typical alternative forms of high value resistors are listed in Table 1. All of these resistors will exhibit noise in excess of that due to thermal fluctuations. This 'current noise' is determined by the basic properties of the resistive material and also by the dimensions of the resistive element. In the case of a film resistor the current noise is lowered if the voltage gradient is reduced by using the maximum length of helical track. This effect is partly offset by an increase in noise if the track becomes too narrow, and thus an optimum condition has to be determined, usually by empirical experimentation. The current noise of the resistor is also reduced if the film is kept as thick as possible. This effect can predominate even though it may necessitate the use of more diluents in an inhomogenous resistive system resulting in an increase in the basic noise-producing tendency of the material. A long resistance track of maximum thickness and reasonable width has been found empirically to produce high value film resistors with minimum current noise.

It is not surprising that solid composition resistors tend to have high noise level when the path length is short and there is a high potential gradient. Deposited carbon film resistors are reasonably low in noise as they are formed from a pure elemental carbon with low basic noise-producing tendency. When high value carbon resistors are produced, however, the films are so thin that the resistors may be more noisy than those made from thicker films of a carbon/resin composition. The high-value metal oxide resistors, even though they include an inert insulating phase as a diluent, exhibit the lowest noise as they combine reasonably low basic noise with an optimized film geometry.

The noise measurements given in Table 1 were made in accord with a proposed international standard specification and the results are expressed in decibels referred to one r.m.s. microvolt of noise per d.c. volt applied to the resistor in a decade of frequency. A maximum of 300V d.c. was applied to the resistors during the noise measurements. As is usually the case, the measurements of noise on nominally identical resistors in a given batch are distributed around the mean values which are quoted in the table. In the case of the high voltage Metox resistors the lowest measurements recorded in a batch at any resistance value coincided with the lowest discernible noise level which could be measured.

These resistors were not developed primarily as low noise units, but the fortuitous discovery that they have a level of noise significantly lower than any other film or solid form of resistor of similar value has led to their finding favour in a number of applications.

REFERENCES

1. KIRBY, P. L., WALTON, B. Recent Developments in Resistors for Use at High Voltages. *Proc. Instn. Elect. Engrs.* *109B*, 47 (1962).
2. British Patent No. 857 400

NATO High Power V.L.F. Radio-Telegraph Station

To augment the communications facilities of the North Atlantic Treaty Organization, the General Post Office has been responsible for building a high-power very low frequency radio-telegraph station at Anthorn on the coast of Cumberland which was brought into operation towards the end of last year.

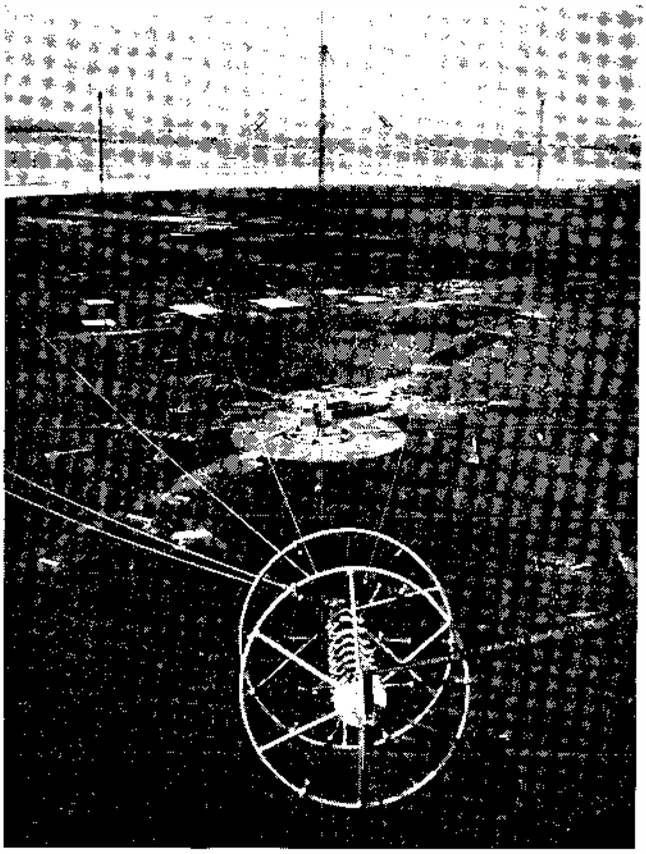
The site, which is a disused airfield between the villages of Anthorn and Cardurnock on the Solway Firth, is particularly suitable for a v.l.f. station; it has an area of some 700 acres at the extremity of a peninsula in a region of fairly flat land of low resistivity, the subsoil is capable of supporting heavy mast loadings without undue difficulty and an adequate public power supply is available.

Aerial System

The aerial system which is claimed to be the largest in Europe consists of six rhombic shaped sections radiating from the top of a 748ft high centre mast. The distant points of the rhombic sections are supported by six masts 618ft high forming an outer ring of about 2150ft, while the mid points of the sections are secured by six masts 678ft high and about 1280ft distant from the centre mast. This arrangement of the aerial enables each section to be lowered to the ground for maintenance without fouling mast stays or any other obstructions. Each section can be raised or lowered by four halyard winches, remotely operated from a control room on the roof of the transmitter building. The aerial has been built of steel-cored aluminium conductors about 1in in diameter to enable it to withstand the working voltage (120kV) without perceptible corona under adverse weather conditions, and also be strong enough to support a heavy coating of ice. Should the ice loading on the conductors increase beyond a prescribed value, halyard tension-limiting gear will cause the winches to lower the aerial automatically so as to prevent the halyard tension rising above a safe maximum value. The aerial is insulated by strings of compression-type insulators so that failure of any unit insulator will not result in collapse of the aerial.

The galvanized steel masts which were designed to withstand wind speeds ranging from 100 mile/h at ground level to 130 mile/h at the top of the highest mast are of conventional design, triangular in section, stayed in three directions at intervals of about 150ft and pivoted at the

The centre mast and transmitter building as seen from the top of one of the 618ft outer masts



base. The posts are of solid round section to minimize wind loading and will, in the worst case, transmit a maximum thrust approaching 1000 tons to the mast foundation. 1500 tons of structural steel and 250 tons of pre-stressed steel wire rope have been used for the masts and stays.

On each mast provision is made for a weatherproof self-climbing passenger hoist operating on the rack and pinion principle. Permanent tracking is fitted to all masts but as the hoist cage is transportable it is not necessary to have a cage for each mast.

The aerial is tunable over the frequency range 16 to 20kc/s by variometers in series with a fixed helical inductor. One second variometer is automatically controlled to compensate for variations in aerial capacitance due to weather effects and so ensure that the aerial circuit is always tuned. These inductors are accommodated in a copper-lined room in the transmitter building and the lead-out to the aerial is taken via a large bushing in the wall of the inductor room.

The contract, valued at £1¼M for the design, supply and erection of the thirteen masts together with the complete aerial system was placed with British Insulated Callender's Construction Co Ltd.

Transmitter and Building

The transmitter building which has a floor space of 20 000ft^2 is at the centre of the aerial system and has been extensively screened against electrical losses. This screening is connected via a mesh of wires to an earth system which consists of about 100 miles of wire laid radially at 2° intervals and at a depth of 1ft.

The transmitter comprises a 250mW drive stage followed by five stages of amplification and, at its normal working frequency of 19kc/s, it is capable of delivering a power of 550kW.

The radio-frequency valves are air-cooled and the mercury vapour h.v. rectifier valves are provided with overload protection equipment which will operate to remove power from the transmitter in about 5μsec. To avoid interruption due to failure of valves or other equipment, the transmitter consists of two 275kW power amplifiers with tuned input and output circuits and which can be worked individually or in parallel.

The tank-circuit for the output stage of each half of the transmitter comprises a fixed inductor of toroidal form connected in parallel with variable oil-filled capacitors. The use of the toroidal form for the inductor minimizes the external field and, therefore, the space required for this item. Coupling to the aerial-circuit is effected via matching networks which enable the aerial to be energized by either or both halves of the transmitter.

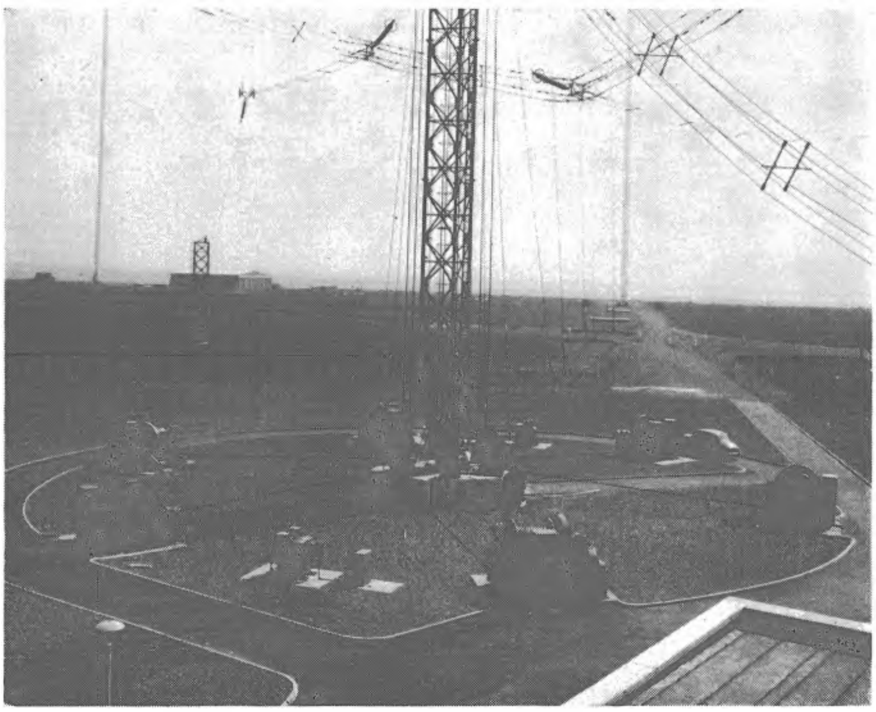
The carrier frequency of the 550kW transmitter is controlled with an accuracy of at least 1 in 10^{10} per day (1 in 10^9 per month) and can be set in steps of 0·1kc/s over the range from 16 to 20kc/s.

Variable frequency oscillators are provided for use in conjunction with the frequency synthesizers to allow any frequency over the operating range to be set with an accuracy of one quarter of a cycle per second. The frequency synthesizer can be set from the control console to any of four preset frequencies in the operating range. These preset frequencies can be changed within a few seconds.

Telegraph signals modulate this carrier in a drive unit, the output from which is applied to the low-power amplifying stages of the transmitter. The whole transmitter and aerial system is capable of operation at signalling speeds up to 50 bauds with frequency-shift or amplitude modulation.

The entire transmitter and drive system is controlled from a console, comprising eight control desks. Four of the desks are used to control the individual stages of the transmitter. A further desk controls the aerial tuning and load system, and another carries a master control panel providing complete control of the system and allowing up to four preset frequencies to be selected, as well as incorporating an illuminated 'transmitter-state' diagram. The other control desks provide monitoring facilities and controls for remote operation, including starting and switching of standby diesel generators.

The arrangement of the six halyard winches around the base of the centre mast



The power supply will normally be taken from the mains but two 600kW generators are provided for standby purposes and these can be started remotely from the console. Each generator is connected so as to supply independently one half of the transmitter and miscellaneous loads.

The main contract, valued at about £3¾M was placed with the Continental Electronics Systems Incorporated of Dallas, Texas, who were responsible for the design and manufacture of the transmitters.

Redifon Ltd of London who are an associate company of Continental Electronics Systems, designed and manufactured the frequency synthesis and drive equipment. The complete control console was manufactured at the Communications Laboratories of Redifon located at Crawley.

The Power Gain of Arrays of Yagi Aerials

By P. Knight*, B.A., A.M.I.E.E., and R. D. C. Thoday*

Broadside arrays of Yagi aerials are frequently used in the v.h.f. and u.h.f. bands in order to obtain increased gain and directivity. The conventional method of gain calculation based on the integration of the power radiation pattern is not easy to apply, however, because of the complicated shapes of the patterns of these arrays. This article describes an alternative method of gain calculation which depends on the evaluation of the mutual resistance between pairs of Yagis. It also contains a table from which the gains of a range of Yagi arrays can be readily determined.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

In order to obtain increased gain and directivity, aerials consisting of arrays of Yagi aerials are commonly used both for transmission and reception in the v.h.f. and u.h.f. bands. A typical arrangement is shown in Fig. 1, which represents a broadcasting aerial having moderate directivity in the horizontal plane and much greater directivity in

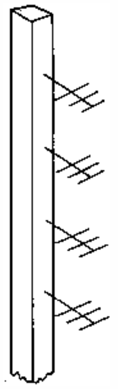


Fig. 1. An array of Yagi aerials on a mast

the vertical plane. However, difficulty arises in calculating the power gain of this type of aerial, because the method previously described[1] for single Yagis cannot be used. Nor can tables[2] for the gains of aerials having approximately omnidirectional horizontal radiation patterns (h.r.p.) be used, since the variation in the h.r.p. of the type of aerial shown in Fig. 1 is too large. Although it is possible to compute the gain of any aerial by integrating the three-dimensional power radiation pattern, the calculation would be very laborious if applied to this type of aerial. The alternative method of gain calculation described in this article was therefore devised to overcome these difficulties. The method enables the gain of the array relative to that of a single Yagi to be calculated, and tables of such gains are included. The gain of the array relative to a half-wave dipole is then obtained by multiplying by the gain of a single Yagi; the latter may be readily determined from its radiation pattern[1]. Although the method applies strictly to arrays of Yagis in free space it may be applied with little error to arrays of the type shown in Fig. 1 provided the support structure does not have an appreciable effect on the radiation pattern.

Method of Gain Calculation

The currents in the individual elements of a three-element Yagi aerial were first estimated from its measured radiation pattern. The relative mutual resistance between a pair of parallel Yagis was then calculated from mutual resistance tables for parallel dipoles. This calculation was repeated for a number of spacings and a table of relative mutual resistances drawn up. Tables of gains of arrays of Yagis were then computed from the relative mutual resistances

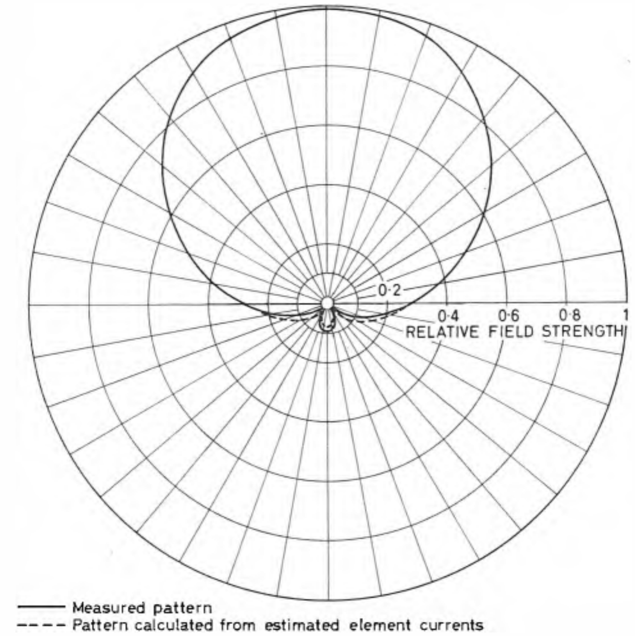


Fig. 2. H-plane patterns of a three-element Yagi

between the individual Yagis. The method is described in detail below.

Estimation of the Currents in the Elements of a Yagi

Fig. 2 shows the measured H-plane pattern of a typical three-element Yagi. In this aerial the spacings between the driven element and director and reflector were 0·13λ and 0·24λ respectively but, so as to simplify the calculation of the relative mutual resistance table, both spacings were assumed to be equal to 0·19λ. In estimating the value of I_D (current in the director) and I_R (current in the reflector) vector diagrams of the distant field components at bearings of 0°, 90° and 180° to the line of the array were at first constructed using known approximate values of parasitic element currents for typical Yagis. The amplitudes and phases of these currents were successively adjusted until the relative fields at a number of bearings were close to the measured values. Particular attention was paid to the ratio between the forward field and that at right-angles to the array, since the coupling between parallel Yagis depends on the radiation at right-angles to their axes. Fig. 2 shows that reasonably good agreement between the measured pattern and that calculated from the estimated currents was obtained; exact agree-

* British Broadcasting Corporation.

ment would not be expected because the spacings between the elements are not the same in the two cases.

The estimated currents* for the equally-spaced Yagi are given in Table 1.

TABLE 1

Estimated currents in three-element Yagi

Current in reflector (I_R)	$0{\cdot}365\angle-146°$
Current in driven element	$1\angle 0°$
Current in director (I_D)	$0{\cdot}63\angle-130°$

It was subsequently confirmed that these currents were consistent with the mutual impedances between the three elements. This was necessary because it may be shown that with three-element array a given radiation pattern may be established by four different current ratios. Only one of these ratios, however, can be obtained in practice if two of the elements are parasitic.

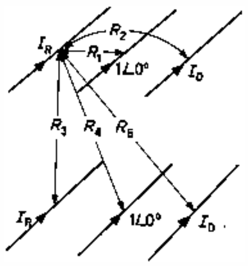


Fig. 3. Two parallel three-element Yagis

Relative Mutual Resistance Between Parallel Yagis

Fig. 3 shows two equally-driven parallel three-element Yagis in which I_R and I_D are the normalized reflector and director currents given in Table 1. It is assumed that these current ratios are not modified by the presence of other Yagis. This may not be strictly true in practice but it is thought that the error introduced by the small changes in current which occur does not introduce any significant error in the gain calculation.

Each of the Yagi elements may be regarded as a driven half-wave dipole and one may proceed to calculate the mutual resistance between Yagis from the mutual resistances between parallel dipoles. If the elements in the Yagis are equally spaced, only the five mutual resistance values shown in Fig. 3 are required.

If R_o is the self-resistance of a single element it may be shown from equation (15) of reference 2 that the power radiated by a single Yagi on its own is:

$$\begin{aligned} W_1 &= R_oI_RI_R^* + R_1I_R + R_2I_RI_D^* + R_1I_R^* + \\ &\qquad R_o + R_1I_D^* + R_2I_DI_R^* + R_1I_D + R_oI_DI_D^* \\ &= R_o\,(1 + |I_R|^2 + |I_D|^2) + 2R_1(\mathrm{Re}I_R + \mathrm{Re}I_D) + \\ &\qquad 2R_2(\mathrm{Re}I_R\mathrm{Re}I_D + \mathrm{Im}I_R\mathrm{Im}I_D) \quad \ldots\ldots\ldots (1) \end{aligned}$$

where I^* indicates the complex conjugate of I, and Re and Im denote Real and Imaginary parts respectively.

Similarly, it may be shown that the power radiated by a single Yagi in the presence of another is:

$$\begin{aligned} W_2 &= W_1 + R_3I_RI_R^* + R_4I_R + R_5I_RI_D^* + R_4I_R^* + \\ &\qquad R_3 + R_4I_D^* + R_5I_DI_R^* + R_4I_D + R_3I_DI_D^* \\ &= W_1 + R_3(1+|I_R|^2+|I_D|^2) + 2R_4(\mathrm{Re}I_R+\mathrm{Re}I_D) + \\ &\qquad 2R_5(\mathrm{Re}I_R\mathrm{Re}I_D + \mathrm{Im}I_R\mathrm{Im}I_D) \quad \ldots\ldots\ldots (2) \end{aligned}$$

Let r be the relative mutual resistance between the Yagis.

* *The elements were assumed to be of equal length but in a practical aerial, because of unequal element lengths, the current at the centre of the reflector would have to be slightly less and that at the centre of the director slightly greater, in order to obtain the same radiation pattern.*

Then:

$$W_2 = W_1(1 + r) \quad \ldots\ldots\ldots\ldots\ldots (3)$$

$$r = \frac{W_2 - W_1}{W_1} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

Substituting the values of W_1 and W_2 given by equations (1) and (2) into equation (3):

$$r = \frac{\alpha R_3 + \beta R_4 + \gamma R_5}{\alpha R_o + \beta R_1 + \gamma R_2} \quad \ldots\ldots\ldots\ldots (5)$$

where $\alpha = 1 + |I_R|^2 + |I_D|^2$

$\beta = 2(\mathrm{Re}I_R + \mathrm{Re}I_D)$

$\gamma = 2(\mathrm{Re}I_R\mathrm{Re}I_D + \mathrm{Im}I_R\mathrm{Im}I_D)$

Values of relative mutual resistance calculated from equation (5) are given in Table 2, the appropriate values of R_1 to R_5 being obtained from tables of mutual impedances between parallel half-wave dipoles. Table 2 contains

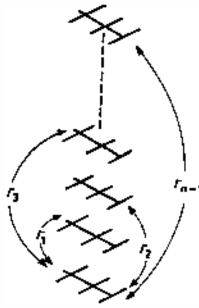


Fig. 4. An array of equally spaced Yagis

all the values of mutual resistance required for gain calculations on arrays of Yagis having uniform spacings in the range 0·5 to 1·20λ.

The Gain of Arrays of Yagis

Fig. 4 shows an array of n equally-spaced Yagis, in which r_q is the relative mutual resistance between two Yagis spaced q times the spacing between adjacent Yagis.

TABLE 2

Relative mutual resistance between parallel three-element Yagis

d/λ	r	d/λ	r	d/λ	r	d/λ	r
0·50	−·117	1·00	−·042	1·50	+·023	2·00	−·006
0·55	−·027	1·10	−·008	1·65	−·034	2·20	−·039
0·60	−·049	1·20	−·049	1·80	−·051	2·40	+·028
0·65	−·105	1·30	+·071	1·95	−·025	2·60	−·014
0·70	−·137	1·40	+·062	2·10	+·020	2·80	−·031
0·75	−·148	1·50	+·023	2·25	+·042	3·00	−·005
0·80	−·152	1·60	−·018	2·40	+·028	3·20	+.024
0·85	−·142	1·70	−·046	2·55	−·004	3·40	+·018
0·90	−·119	1·80	−·051	2·70	−·028	3·60	−·011
0·95	−·078	1·90	−·039	2·85	−·028	3·80	−·023
1·00	−·042	2·00	−·006	3·00	−·005	4·00	−·003
1·05	−·016	2·10	−·020	3·15	+·018	4·20	+·018
1·10	+·008	2·20	−·039	3·30	+·027	4·40	+·014
1·15	+·029	2·30	−·042	3·45	+·011	4·60	−·009
1·20	+·049	2·40	−·028	3·60	−·011	4·80	−·018
d/λ	r	d/λ	r	d/λ	r	d/λ	r
2·50	−·007	3·00	−·005	3·50	+·004	4·00	−·003
2·75	−·031	3·30	+·027	3·85	−·021	4·40	+·014
3·00	−·005	3·60	−·011	4·20	+·018	4·80	−·018
3·25	+·027	3·90	−·015	4·55	−·003	5.20	+·015
3·50	−·004	4·20	−·018	4·90	−·010	5·60	−·008
3·75	−·023	4·50	+·003	5·25	−·016	6·00	−·001
4·00	−·003	4·80	−·018	5·60	−·008	6·40	+·009
4·25	+·020	5·10	−·009	5·95	−·004	6·80	−·012
4·50	+·003	5·40	−·011	6·30	−·013	7·20	+·011
4·75	−·019	5·70	−·015	6·65	−·009	7·60	−·006
5·00	+·000	6·00	−·001	7·00	−·001	8·00	−·001
5·25	+·016	6·30	−·013	7·35	+·011	8·40	−·006
5·50	−·002	6·60	−·006	7·70	−·010	8·80	−·010
5·75	−·015	6·90	−·007	8·05	+·003	9·20	+·008
6·00	+·001	7·20	−·011	8·40	+·006	9·60	−·005

For an isolated Yagi radiating unit field, the radiated power is W_1. For n Yagis carrying the same currents, the radiated field is n times that radiated by a single Yagi. If the Yagis are equally spaced it may be shown that the power radiated by the array is:

$$W_n = W_1 \; [n + 2\{(n-1)r_1 + (n-2)r_2 + \ldots + r_{n-1}\}]$$

$$= W_1 [n + 2\sum_{q=1}^{n-1} (n-q)r_q] \quad \ldots\ldots\ldots\ldots\ldots \quad (6)$$

The gain of n Yagis relative to a single Yagi is therefore:

$$G_n = n^2W_1/W_n = \frac{n^2}{n + 2\sum_{q=1}^{n-1}(n-q)r_q} \quad \ldots\ldots \quad (7)$$

It is usually more convenient to consider the gain per Yagi since this is a quantity which varies slowly with the number of Yagis and which tends to a limiting value as the number of Yagis increases. From equation (7) the expression for the gain per Yagi is seen to be

$$G = \frac{n}{n + 2\sum_{q=1}^{n-1}(n-q)r_q} \quad \ldots\ldots\ldots\ldots \quad (8)$$

Gains for arrays of up to eight Yagis having spacings in the range 0·50λ to 1·20λ were calculated from equation (8) on a digital computer and are contained in Table 3. It should be noted that this table gives the gain of an array relative to that of one of the Yagis comprising it. To determine the gain of the array relative to a standard aerial such as λ/2 dipole it is necessary to multiply this figure by the gain of a single Yagi*. The figure thus derived for the array gain applies to the direction in which the Yagis point, which is usually the direction in which the effective radiated power is greatest. It should not be confused with the mean gain, which may be several decibels lower.

The gain calculation may be extended to unequally spaced arrays, or to arrays carrying unequal currents, by substituting relative mutual resistance values from Table 2 into the appropriate equation from Section 2·5 of reference 2.

The method may also be used if each level or tier of an aerial contains two or more Yagis pointing in different directions, as shown in Fig. 5; the aerial then consists of a number of columns of Yagis spaced around the mast. If the mutual impedances between the columns can be neglected, as is usually the case, each column may be treated as a separate array and Table 3 used to assess the gain in the following manner.

From equation (7) the power radiated by one column is:

$$W_n = n^2W_1/G_n = \frac{n^2K\,|I|^2}{G_n} \quad \ldots\ldots\ldots\ldots \quad (9)$$

where I is the current fed to each Yagi in the column and K is a constant. G_n is derived from Table 3 by multiplying by n. Suppose the aerial has three identical columns whose individual Yagis are fed with currents I_A, I_B and I_C. Applying the result given in equation (9), the total power fed to the array of $3n$ Yagis is:

$$W_{3n} = n^2K\,(|I_A|^2 + |I_B|^2 + |I_C|^2)/G_n \quad \ldots\ldots \quad (10)$$

The field E_1 radiated by a single Yagi in the direction in which it points is proportional to its feed current and may

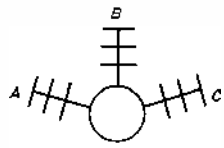


Fig. 5. An aerial containing three columns of Yagis

be written in the form $E_1 = sI$ where s is a constant. The direction in which the gain of the aerial is to be calculated must now be specified; assume it is in the direction in which the Yagis of column A point. The contributions in this direction from the other two columns of Yagis cannot be disregarded; thus the field in this direction is the vector sum of contributions from all three columns and is given by:

$$E_{3n} = ns(I_A + pI_B + qI_C) \quad \ldots\ldots\ldots\ldots \quad (11)$$

where p and q are complex constants, determined by the h.r.p. of the individual columns. The gain of the complete aerial is therefore:

$$G_{3n} = (W_1/W_{3n})\,|E_{3n}/E_1|^2 = G_n\,\frac{|I_A + pI_B + qI_C|^2}{|I_A|^2 + |I_B|^2 + |I_C|^2} \quad \ldots \quad (12)$$

since $W_1 = k|I|^2$ and $E_1 = sI$. This expression gives the gain in the direction in which column A points; the gain in other directions can, of course, be determined from the h.r.p.

Application of the Method to Other Arrangements

Tables 2 and 3 may be applied to Yagis having different numbers of radiating elements or different spacings provided their H-plane patterns do not differ appreciably from those shown in Fig. 3. If, however, the patterns are very different, further calculations along the lines described would be required. A similar method could be used for arrays of collinear Yagis*, but little would be gained by doing this because the mutual impedances between collinear Yagis are much smaller than those between parallel Yagis; the gains of arrays consisting of more than 2 collinear Yagis are therefore largely independent of their spacing.

Acknowledgments

Thanks are due to Mr. R. W. Lee for writing the computer program which enabled Table 3 to be computed, and to the Director of Engineering of the BBC for permission to publish the article.

TABLE 3

Gain per Yagi of parallel three-element Yagis

SPACING (λ)	NUMBER OF YAGIS 2	3	4	5	6	7	8
0·50	0·895	0·887	0·873	0·867	0·862	0·859	0·856
0·55	0·974	0·960	0·969	0·960	0·964	0·959	0·961
0·60	1·052	1·034	1·053	1·052	1·053	1·057	1·056
0·65	1·117	1·102	1·110	1·122	1·118	1·121	1·124
0·70	1·159	1·165	1·154	1·164	1·170	1·166	1·167
0·75	1·174	1·222	1·217	1·216	1·227	1·234	1·233
0·80	1·179	1·273	1·302	1·303	1·306	1·317	1·328
0·85	1·166	1·282	1·353	1·386	1·395	1·397	1·401
0·90	1·135	1·239	1·322	1·386	1·430	1·457	1·470
0·95	1·085	1·149	1·205	1·255	1·302	1·345	1·383
1·00	1·044	1·064	1·077	1·086	1·093	1·097	1·101
1·05	1·016	1·008	0·995	0·980	0·966	0·952	0·940
1·10	0·992	0·965	0·939	0·920	0·907	0·900	0·896
1·15	0·972	0·938	0·917	0·907	0·906	0·906	0·905
1·20	0·953	0·923	0·912	0·912	0·912	0·909	0·906

Note: This table gives the gain of the array relative to that of a single Yagi in free space.

REFERENCES

1. MANTON, R. G. The Calculation and Measurement of the Gains of End-Fire V.H.F. and U.H.F. Aerials. *Electronic Engng.* 36, 8 (1964).
2. KNIGHT, P., MONTEATH, G. D. The Power Gain of Multi-Tiered V.H.F. Transmitting Aerials. BBC Engineering Division Monograph No. 31 (July 1960).

* *The gain of the three-element Yagi described is 3·7 (5·7dB). The gains of Yagis having slightly different radiation patterns may be calculated by the method described in reference 1.*

* *Yagis arranged side by side with their elements in the same plane. The arrangement shown in Fig. 1 with the Yagis rotated to transmit vertical polarization, would constitute such an array.*

A Review of Methods of

Post Deflexion Acceleration Used in Instrument Cathode-Ray Tubes

By K. J. Ming*, B.Sc.

An outline is given of the operation of cathode-ray tubes which are commonly used in electronic instruments at the present time. These tubes mostly incorporate some form of post deflexion acceleration, either the simple band type or a graphite spiral.

Tubes with greater deflexion sensitivity are now becoming available as a result of improved post deflexion acceleration systems which employ a grid or mesh electrode. The characteristic features of three such systems, namely the axial accelerator, the radial accelerator and the divergent lens accelerator, are described and compared.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

ONE of the most useful and frequently used pieces of electronic equipment is the cathode-ray oscilloscope. The specifications of these instruments have improved gradually, partly due to advances in circuit techniques, but also because better cathode-ray tubes have become available.

In the simplest form of cathode-ray tube, electrons emitted from the cathode are accelerated by positive potentials in a suitable gun structure to form a spot on the fluorescent screen. Deflexion of the spot is produced by two pairs of deflector plates placed mutually at right-angles and mounted in such a position that the electron beam from the gun passes successively between the two plates of each pair. In such a tube, if the voltage on the final accelerating anode is increased to produce a brighter spot, a proportionately larger deflexion voltage is required, due to the increased energy of the electrons.

However, if the system is modified to enable the beam to be deflected while the electrons are travelling at low velocity and subsequently the electrons are further accelerated by an anode at a higher voltage, a high brightness can be achieved while retaining a high deflexion sensitivity. This arrangement is known as post deflexion acceleration (p.d.a.) and is used in most instrument tubes in present use.

Band P.D.A.

The earliest form of p.d.a. is shown in **Fig. 1**. The internal graphite coating on the bulb is divided into two sections by an insulating band in the flared part of the bulb. In operation the band of graphite nearer the electron gun is maintained at the potential of the final gun anode and an accelerating voltage is connected between the two sections of the graphite. As the accelerating voltage is increased the display becomes brighter. In addition the size of the display is reduced and distortion introduced, which can be seen as barrel distortion of a rectangular raster.

The magnitude of the acceleration employed is usually expressed as the ratio of the final accelerating voltage to the final gun anode voltage (p.d.a. ratio) and the amount of shrinkage of the scan by scan efficiency, defined as

$$\frac{\text{scan size with p.d.a.}}{\text{scan size without p.d.a.}} \times 100 \text{ per cent}$$

The relation between scan efficiency and p.d.a. ratio for a band p.d.a. can be seen in Fig. 2.

This type of p.d.a. arrangement is, in effect, a simple bi-potential lens. A sketch showing the equipotential surfaces of the electrostatic field is shown in Fig. 3. This is a convergent lens which is stronger at the periphery than at the centre. The p.d.a. ratio employed is limited by shrinkage of the raster and degree of distortion which can

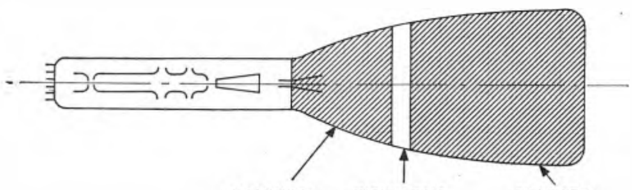
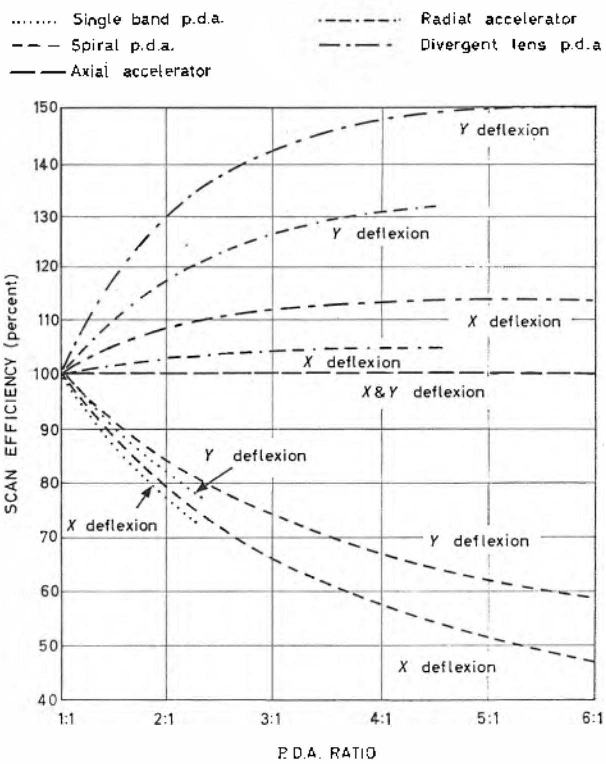


Fig. 1 (above). Band type accelerator

Fig. 2 (below). Variation of scan efficiency with p.d.a. ratio for various types of p.d.a.



* *The M-O Valve Co. Ltd.*

be tolerated. In practice a p.d.a. ratio of about 2:1 is used.

A way of reducing the undesirable effects of the lens is by the use of several insulating bands in the graphite coating. This allows successively higher accelerating voltages to be applied to the sections of graphite, which results in a series of weak lenses.

Spiral P.D.A.

The simple band types of p.d.a. described in the previous section have largely been superseded by a distributed accelerator system. A sketch of this system is shown in Fig. 4. A graphite spiral is painted on the inside of the bulb extending from the region of the deflector plates nearly to the screen. The spiral is terminated at both ends by graphite bands. The band at the neck end is connected to the final anode of the electron gun, and the accelerating potential is connected across the spiral. In many cases, the fluorescent screen is aluminized, which provides a con-

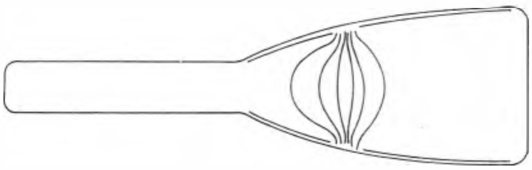


Fig. 3. The lens in the band p.d.a. showing the equipotential surfaces

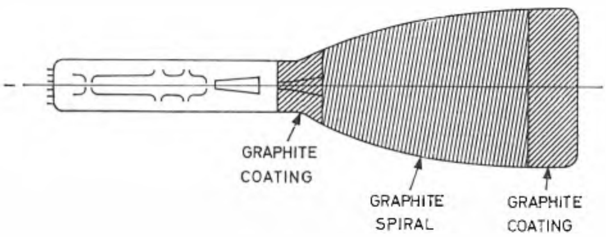


Fig. 4. Spiral p.d.a.

tinuous conducting coating to the bulb at that end, while in non-aluminized tubes the secondary emission characteristics of the screen are such that its potential is maintained at that of the surrounding graphite band.

The electrostatic field of a spiral p.d.a. can be considered to comprise three sections:

(1) A cylindrical lens, which is convergent in the direction of deflexion of the front plates and is formed by the penetration of the p.d.a. field into the front deflector plate region.

(2) A region in which the equipotential surfaces are parallel planes orthogonal to the axis of the tube which will cause a reduction of the deflexion of a deflected beam.

(3) An axially symmetric divergent lens section at the screen end where the p.d.a. field penetrates into the region bounded by the cylindrical graphited walls of the bulb and the screen.

The relation between the scan efficiency and p.d.a. ratio is shown in Fig. 2. With spiral p.d.a. systems of this sort, p.d.a. ratios up to about 6:1 have been used to give acceptable results regarding scan distortion. At 6:1 p.d.a. ratio the scan efficiencies are about 45 per cent in the X direction and 55 per cent in the Y direction.

A further effect of the types of p.d.a. already described is that the spot size is reduced. However, this is counterbalanced to some extent by the increase in deflexion defocusing due to the increased angle through which the electron beam has to be deflected.

High Efficiency P.D.A. Systems

Recently p.d.a. systems have been developed which have an efficiency in the region of 100 per cent. The 'ideal' p.d.a. system would be a system which would accelerate the electrons along their paths only, thereby increasing the brightness of the display without having any other effect.

The Axial Accelerator

A system which would achieve this result was put forward by L. S. Allard[1] in 1950. In this system, shown in Fig. 5, the flared part of the bulb is made substantially a field free region by coating the inside of the bulb with graphite and mounting a wire mesh near and parallel to the fluorescent screen. The fluorescent screen has an aluminium coating which is insulated from the graphite on the bulb. This allows a potential to be applied between the mesh, which is maintained at approximately graphite potential, and the screen. An accelerating field is thus produced with lines of force parallel to the axis of the tube, i.e. the arrangement is an axial accelerator. As the

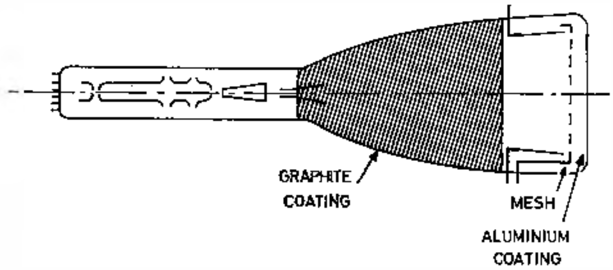


Fig. 5. Axial accelerator

distance over which the field acts is small, its action is very near to an acceleration along the paths of the electrons.

In practice the mesh is supported on a metal truncated cone to allow a greater gap between the graphite and the screen, to reduce the possibility of high voltage breakdown over the surface of the glass.

With such a p.d.a. system, when the electron beam strikes the mesh, secondary electrons would be liberated which would be focused by the p.d.a. field into a secondary spot. This could easily be seen, when the beam is deflected, as a diffuse spot a little nearer to the centre of the tube than the primary spot. It has been shown by Schlesinger[2] that by coating the mesh on the gun side with an insulator of high secondary emission coefficient, the field set-up between the back and front sides of the mesh is sufficient to eliminate the effects of the secondary electrons when the mesh is biased negatively by about 100V with respect to the cone.

Tubes incorporating a coated mesh axial accelerator p.d.a. have been made and demonstrated. As predicted, no change in scan size takes place as the p.d.a. ratio is increased, and the upper limit of the p.d.a. ratio is determined only by the high voltage breakdown performance of the arrangement. No scan distortion is introduced by this system of p.d.a. The interstitial lenses formed in the mesh do not produce any observable increase in spot size, although precise measurement is difficult because of the change in brightness level as the p.d.a. ratio is increased. One important feature of the axial accelerator p.d.a. is that there is no reduction of spot size due to the p.d.a. action compared with the spiral p.d.a. Thus if the same electron gun were used in conjunction with the two p.d.a. systems, a larger spot would be produced with the axial accelerator, and hence electron optical modifications are required in the gun to maintain resolution. However as the deflexion angle is less with the axial accelerator, the focus uniformity is improved.

It is necessary to choose the pitch of the mesh to be small enough so that the mesh structure cannot be seen with the unaided eye. Also, since the mesh intercepts part of the beam, an additional requirement is that the electron transmission of the mesh should be as high as possible. Although it has been claimed that in such cases the electron transmission is greater than the optical transmission, the latest information by Donlan and Nicklas[3] indicates that any such effect is negligible. It has been shown that a mesh with about 200 wires per inch and an optical transmission of 70 per cent is satisfactory.

The Radial Accelerator

If tubes have a common centre of deflexion for both X and Y axes, such as with electromagnetic deflexion or by use of the deflectron[2] the 'ideal' condition can be fulfilled by establishing an electrostatic field with a series of equipotential surfaces which are the caps of spheres centred on the deflexion centre. With such a field, the beam always lies along the radius of the spheres. However, in normal instrument tubes where the X and Y deflectors are at different positions on the axis of the tube, it is impossible to attain the 'ideal' condition exactly. Nevertheless the approximation is sufficiently close for practical tubes to be made incorporating a p.d.a. working on these principles. A sketch of such a tube is shown in Fig. 6. A variable pitch graphite spiral is painted on the inside of the bulb, with close turns at the neck end and widely spaced turns at the screen end. The accelerating voltage is connected across the spiral which provides the correct potential distribution in the flared part of the bulb to give spherical equipotential surfaces, i.e. the potential along the radius of the spheres is proportional to the reciprocal of the distance from the centre of the spheres. It is necessary to terminate the field at the entrance by a mesh which is convex and has the correct radius of curvature for its position in the field. In theory the screen should also be curved, but the correct radius of curvature is sufficiently large for a flat screen to be acceptable, especially as the electrons are travelling at high velocity in this region, and are not affected by slight variation from the perfect case.

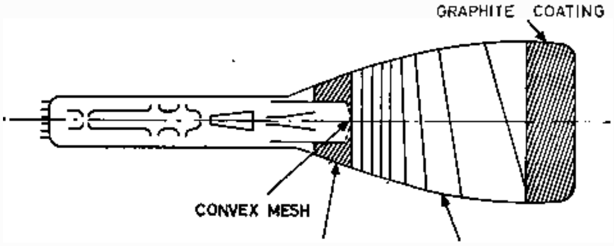


Fig. 6. Radial accelerator

When a centre of deflexion is at a greater distance from the screen than the centre of curvature of the equipotential surfaces, a deflected beam enters the field at an angle to the radius, resulting in an increase in the deflexion due to the p.d.a. field. This scan magnification results in scan efficiencies greater than 100 per cent. The relation between scan efficiency and p.d.a. ratio is shown in Fig. 2. The tube can be operated with a p.d.a. ratio of 4:1, and in this condition the scan efficiencies are 105 per cent and 130 per cent for X and Y directions respectively. While the change from the 'ideal' condition is useful in increasing the deflexion sensitivity, the reduction in the accuracy of the geometry of the p.d.a. system is sufficient to introduce scan distortion which limits the maximum p.d.a. ratio that can be used. Steps can be taken to eliminate this distortion for any particular set of conditions in which the tube is normally used.

Similar distortion can be seen in photographs of a raster on a tube demonstrated by Schlesinger[4] which employs a similar p.d.a. system with a graded spiral and a plane mesh. Although there is substantially no change in the size of the raster up to a p.d.a. ratio of 2½ : 1, noticeable pincushion distortion is seen at a ratio of 3:1 which becomes worse as the ratio is increased.

The radial accelerator p.d.a system, in fact, constitutes a thick divergent electrostatic lens. The term 'divergent lens' in this context has the same meaning as in light optics, i.e. a lens which causes divergence of a path of a beam originally travelling parallel to the axis of the lens.

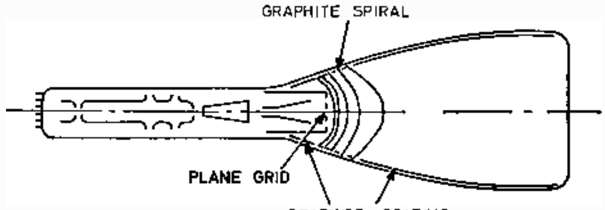


Fig. 7. Divergent lens p.d.a. showing equipotential surfaces

Due to the action of the lens, there is a defocusing effect on the beam which causes an increase in spot size. It is possible that there is also some degradation of the spot size resulting from the interstitial lens action of the mesh. However, as both effects occur simultaneously it is not possible to observe the effect of the latter separately.

When the tube is correctly focused the structure of the mesh cannot be observed in the spot. However, if the focusing is incorrect, the pattern of the mesh can be seen on the screen as a shadow. By making the mesh sufficiently fine (600-800 wires per inch) the pattern is no longer troublesome.

Under conditions of high brightness the effect of secondary emission from the mesh is seen as a diffuse illumination of the screen.

Divergent Lens Accelerator

Instrument tubes incorporating a third form of high efficiency p.d.a. have been available for some years. These tubes incorporate a p.d.a. which is effectively a thin divergent electrostatic lens, i.e. the distance over which the major part of the potential drop takes place along the axis is short compared to the focal length. The arrangement of such a p.d.a. system is shown in Fig. 7.

A fine wire grid is mounted on the end of a metal cylinder which encloses the front deflector plates. The cylinder is connected electrically to the internal graphite coating on the wall of the bulb in the vicinity of the electron gun and is maintained at approximately final gun anode potential. The function of the cylinder is twofold, namely to form one electrode of the lens and also to shield the deflector plates from the field due to the p.d.a. The accelerating potential is connected to the graphite coating on the front part of the bulb, which together with the aluminium coating on the screen forms a substantially field-free region in this part of the bulb. There is a short length of graphite spiral of constant pitch, connecting the two parts of the graphite, which serves to stabilize the potential on the walls of the bulb. The position of the grid is such that it is approximately midway along the length of the spiral.

The potential distribution of the system showing equipotential surfaces is shown in Fig. 7. Provided the lens is strong enough (i.e. a short enough focal length) and far enough from the centres of deflexion, it can provide scan magnification. Variation of scan efficiency with p.d.a. ratio is shown in Fig. 2. It can be seen from this graph that scan efficiencies of 113 per cent and 150 per cent can be achieved in X and Y directions respectively with a p.d.a. ratio of 6:1. The scan distortion at 7:1 ratio is still less than 1 per cent.

As in the case of the radial accelerator the lens action of the p.d.a. results in an increase in spot size. Also the structure of the grid is not seen in the spot provided it is correctly focused. A grid with about 800 wires per inch is usually used to overcome patterning under slightly defocused conditions.

The effect of secondary emission from the grid is noticeable at high brightness level as a diffuse illumination of the screen. This can be substantially reduced by insulating the p.d.a. grid from the supporting cylinder and arranging the cylinder to be slightly positive with respect to the grid. The cylinder then acts as a collector of low voltage secondary electrons from the grid. It is claimed by H. B. Law[5] who used a similar p.d.a. system on a television tube that the effect of secondary emission can be still further reduced by the use of an electroformed mesh instead of a grid. With a mesh which has filaments with a knife-edge cross section, the mesh can be arranged so that the beam does not strike the sides of the apertures as it passes through.

Conclusions

Tubes incorporating some form of high efficiency p.d.a. employing a mesh or grid clearly offer an improvement in deflexion sensitivity over conventional types of p.d.a. The axial accelerator p.d.a. by virtue of its operation cannot introduce scan distortion, and the p.d.a. systems employing lenses give no distortion provided they are used in their correct working conditions.

The variations of spot size produced by various forms of p.d.a. is counterbalanced to some extent by the corresponding variation in deflexion defocusing resulting from the change in deflexion angle. Steps can be taken with any particular design to reduce the spot size by modifications to the electron gun. These modifications, however, would increase the cathode loading which would reduce the effective life of the cathode.

The use of a mesh or grid through which the beam passes can, in some circumstances, produce a corresponding pattern in the spot, but usually this cannot readily be seen, and is not troublesome.

One inevitable feature of using a mesh or grid is the problem of secondary emission. Unless precautions are taken, the secondary emission resulting from an axial accelerator forms an intrinsic secondary spot, while in the case of the divergent lens p.d.a. a general diffuse illumination of the screen is caused. Both of these defects are overcome to varying degrees in currently available tubes either by coating the mesh of the axial accelerator with an insulating coating, or by electrically separating the grid from its support in the divergent lens system and applying appropriate biasing potentials in the two cases.

In addition the interception of part of the beam by the mesh or grid reduces the overall efficiency of the emission system. To reduce this effect the p.d.a. electrode needs to have as high a transmission as possible. This is more readily obtained with the axial accelerator than the lens system because it is possible to use a fairly coarse mesh with a high optical transmission.

Thus, although the spiral p.d.a. tubes which are commonly used today are satisfactory for most purposes and will continue to be used for many years, tubes incorporating some form of high efficiency p.d.a. can be designed to give an improved deflexion sensitivity. Some of these tubes are already available and are beginning to come into use for special requirements. It is possible that in the course of time, as techniques of manufacture improve, and some of the disadvantages are overcome, the use of high efficiency p.d.a. tubes will become more widespread.

REFERENCES

1. ALLARD, L. S. An "Ideal" Post Deflexion Accelerator C.R.T. *Electronic Engng. 22,* 461 (1950).
2. SCHLESINGER, K. Progress in the Development of Post-Acceleration and Electrostatic Deflexion. *Proc. Inst. Radio Engrs. 44,* 659 (1956).
3. DONLAN, P. J., NIKLAS, W. F. Electron Transmission of Mesh Lenses for Scan Magnification in Television Picture Tubes. *J. Brit. Inst. Radio Engrs. 20,* 911 (1960).
4. SCHLESINGER, K. Electron Trigonometry—A New Tool for Electron-Optical Design. *Proc. Inst. Radio Engrs. 49,* 1538 (1961).
5. LAW, H. B., DAVNE, L., RAMBERG, E. G. The Enhanced-Scan, Post-Acceleration Kinescope. *RCA Rev. 22,* 603 (1961).

A Transistorized High-Speed Switch

A new transistorized high-speed switch, requiring absolutely no 'setting-up' or 'adjustment' throughout its entire life, has been developed by The Marconi Co Ltd. This new solid-state switch has been designed to drive electromechanical equipment and may be used as a plug-in replacement for the high-speed telegraph relays used in a wide range of telecommunications equipment today. Over 580 of these switches have already been ordered by the Ministry of Aviation, who will use them to replace telegraph relays in mobile communications equipment.

This robust and reliable switch has many advantages over its mechanical counterpart that requires initial setting-up and regular maintenance. There are no moving contacts and therefore no sparking and corrosion of contact faces. Severe humidity tests have proved that the switch will operate under heavy condensation conditions without any ill effects. Temperature variations from $-10°C$ to $+70°C$ and power supply variations of ± 10 per cent also produced no significant effect on the performance.

The switch has a glass-fibre cover and all connexions are wired to a twelve pin base of the type used by existing high-speed mechanical relays. There are two switches in each container and the inputs and outputs are electrically isolated from each other and from earth. This allows the switch to be connected as two independent switches for separate circuits, or to drive two loads from a common input, or to connect both switches in push-pull to give a single pole changeover action. In the latter configuration the switch can be used as a telegraph relay and when connected to a standard supply, up to twelve 20mA teleprinters can be driven via suitable current limiting resistances from one switch.

The input impedance is 10kΩ per section and two sections may be connected in series or parallel. The input current sensitivity is 0·5mA per section. The maximum output current is 400mA into a short-circuit for 10min at an ambient temperature of less than 30°C. The isolation between input and output is 50MΩ. The device measures $3\frac{1}{4}$in × $3\frac{1}{4}$in × $1\frac{3}{4}$in.

This new device, which is being incorporated in the Company's latest range of equipment is now available from the Specialized Components Division of The Marconi Co Ltd.

A Graphic-Analogue to Digital Convertor

By R. L. G. Gilbert* and N. S. Oakey*

The equipment described in this article will automatically read strip charts, made by the hot stylus process, and present the information in digital form. The geometry of the scanning system is similar to that of a flying-spot scanner and the position of the trace is measured by a counter-timer. The output can be recorded by any digital recorder, e.g. a line printer. The accuracy is ±0·5 per cent f.s.d. and the repeatability ±0·1 per cent.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

ONE of the most common devices used to store data is a strip chart recorder, which gives an analogue record in the form of a line on a strip of paper. This type of recorder is simple, reliable, and can be left unattended for long periods of time. A visual inspection of the record can yield significant information about the experiment and whether or not the measuring and recording system is functioning properly. Detailed analysis of the strip chart record often involves tedious manual reading and subsequent numerical calculation. The device described here was designed to read strip charts automatically and to present the information in digital form.

General Description of the Analogue to Digital Convertor

The equipment was built to read records made with a hot stylus on waxed paper, giving a transparent trace on an opaque background. The geometry of the scanning system is similar to that of a flying spot scanner, with a cathode-ray tube, lens, strip chart, and photomultiplier on a common optical axis. Fig. 1 is a block diagram of the graphic-analogue to digital convertor. A cathode-ray tube provides a spot of light which can be moved across its face. This spot is focused on the paper by means of a lens, and a photomultiplier, mounted directly behind the record, measures the light transmitted through the paper.

Upon receiving a 'start' pulse the sweep generator produces a voltage which rises linearly with time. This is amplified with the deflexion amplifier and fed to the c.r.t. to deflect the spot linearly with time. The image of the spot then starts from near one edge of the paper and moves with constant speed across the strip chart record; as it crosses the trace the photomultiplier will receive a higher light intensity causing its current output to increase. The 'start' signal also triggers a multivibrator which starts a counter-timer, and the increased output from the photomultiplier, when the spot crosses the trace, is used to trigger the multivibrator again and turn off the counter-timer. The counter-timer then registers the time taken for the light spot to move from the edge of the record to the trace. Since the spot moves linearly with time, the timer reading is proportional to the position of the trace.

The timer reading may be recorded with any convenient digital recorder, such as a line printer or digital magnetic tape recorder. Hence, a series of numbers which are proportional to the position of the trace, may be recorded as the strip chart moves, with constant speed, perpendicular to the direction of spot travel across the paper.

The Circuits

The 'start' pulse may be derived from any convenient source depending on the particular recording system used, but in the prototype was obtained from an astable multivibrator. The sweep generator uses a flip-flop to provide a gating waveform for a transistor Bootstrap ramp-voltage generator (which generates a ramp of 1msec duration). The output is coupled to a tube paraphase amplifier whose outputs are connected to the deflexion plates of the c.r.t.

The c.r.t. used is a 3in flat faced tube with a P16 phosphor, the light output of which decays rapidly. The photomultiplier used has a 3in diameter face. Its output (as illustrated in Fig. 2) consists of a 'background' current caused by light passing through the semi-opaque paper and, superimposed on this, a spike which corresponds to the time when the light spot crosses the trace. The output is differentiated to remove the 'background', and the 'spike' is used to trigger a monostable multivibrator which controls the counter-timer gate.

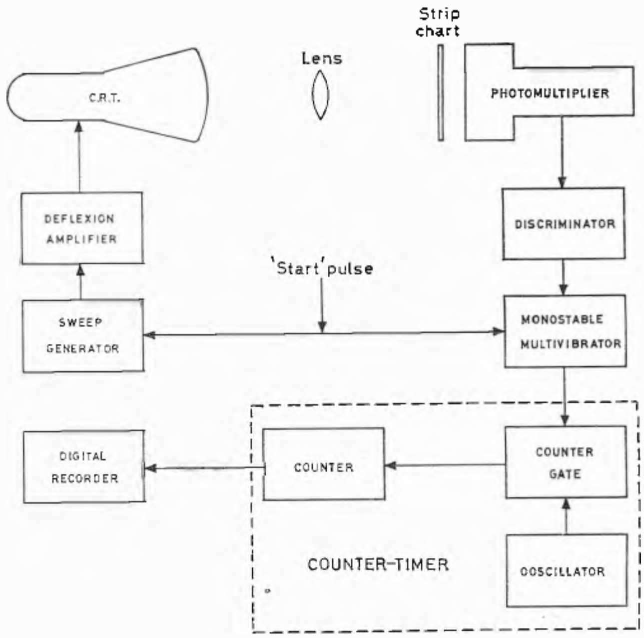


Fig. 1. The graphic-analogue to digital convertor

The counter-timer used is a frequency meter with an internal 1Mc/s oscillator and a series of decade counters whose input is controlled by a gate. When a permanent record of the counter-timer readings is desired they are printed with a digital printer, but any suitable type of recorder can be used.

The Accuracy of the Analogue to Digital Convertor

The accuracy desired was 1 per cent f.s.d. or better, with a repeatability of 0·1 per cent. In designing the device, the trace width, the c.r.t. spot diameter, the c.r.t. phosphor characteristics, the rise time of the electronic circuit, the linearity of the sawtooth voltage, background

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and noise in the photomultiplier output, and the quality of the optical system must be considered. The strip chart paper has a usable width of 50mm and a typical trace width of $\frac{1}{4}$mm. Since the trace has a finite width and one is interested in the position of its centre it is desirable to have a photomultiplier output pulse whose time width is no larger than 1 per cent of the total scan to guarantee 1 per cent accuracy in the reading. With a projected spot diameter on the paper of $\frac{1}{4}$mm the sum of the trace width and spot diameter would be $\frac{1}{2}$mm (Fig. 3) which gives a photomultiplier output whose width is less than 1 per cent of the total scan (using a c.r.t. with rapidly decaying phosphor). The need for linearity of the sawtooth voltage from the sweep generator is obvious, and the rise time of the electronic circuits must be small compared with the time it takes the spot of light to move across the trace to give a good signal-to-noise ratio. The optical system must have good flatness of field to preserve the focus of the spot image on the paper over its width.

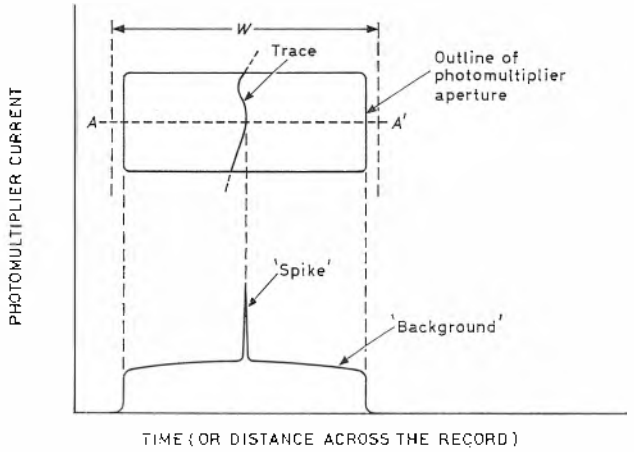
To have a repeatability of 0·1 per cent the sawtooth rate of rise must be very constant; the photomultiplier output must be free of amplitude variations between successive sweeps to assure that the multivibrator triggers at the same point for successive scans of the same trace positions, and there must be no jitter on the c.r.t. spot. A further requirement is that the total sweep time of the spot be such that the timer can measure it to 0·1 per cent accuracy, which implies that the counter-timer must display at least 1 000 counts.

Results

A record was produced with successive steps whose height increased by approximately 5mm each step. This record was run through the graphic-analogue to digital convertor to determine its linearity and repeatability. The record was read with an accuracy of $\pm 0{\cdot}1$mm manually and from these readings and the convertor readings a line was calculated by the method of least squares, giving $Y = 20{\cdot}36X + 53{\cdot}9$ where Y is the number of counts recorded from the counter timer and X is the distance across the record. By calculating Y for various positions and comparing these calculated values with measured values one can determine the linearity of the instrument. The measured values of X and Y, the calculated values of

Fig. 2. The photomultiplier current is shown below as a function of time (or distance across the record). Above is indicated the record as seen by the photomultiplier

The photomultiplier aperture is indicated by a heavy rectangle, the strip chart width is W, and the light spot moves from A to A′. When the spot comes to the edge of the aperture the light through the record causes the 'background' and when the spot crosses the trace the higher intensity of light transmitted causes the 'spike'. At the other side of the aperture the 'background' again becomes zero



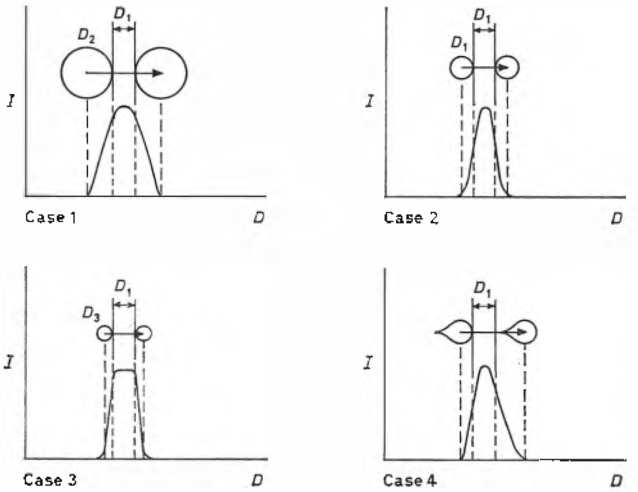


Fig. 3. Illustration of the effective trace width with an actual trace width D_1 and various sizes of light spot

The ordinate is photomultiplier current and the abscissa is the distance across the record the spot has moved. The curves show the amount of light transmitted (which is proportional to the photomultiplier current) as the spot moves across the portion of the record near the trace. Considering the spot moving with constant speed, the heavy trace also indicates the current versus time output of the photomultiplier. In cases 1, 2, and 3 a simple light spot is considered, while in case 4 the shape is meant to illustrate the spot on a c.r.t. whose intensity decays exponentially. This has the same effect as a spot of uniform intensity whose size decreases at one end. The length of the spot in a particular case will depend upon its speed and the decay time of the phosphor. This lengthened spot gives the output pulse a lengthened fall time

Y and the differences and percentage differences are shown in Table 1 and the percentage differences are plotted in Fig. 4. It is seen that the largest difference is $+0{\cdot}4$ per cent, the overall linearity of the instrument is better than $\pm 0{\cdot}5$ per cent f.s.d.

To determine the repeatability, a trace was read many times at the same place over a period of many minutes. The answers never differed by more than ± 1 count so the repeatability is $\pm 0{\cdot}1$ per cent f.s.d.

In another test a series of lines of constant slope were drawn on the record and digitized. The constancy of the differences between readings is a measure of the linearity of the device. A sample of this test is shown in Table 2.

A further test was to read a record of waves. This was done and the values were plotted and are shown on Fig. 5 compared with the corresponding section of the actual record.

Future Developments

A scale of two flip-flop will be triggered from the photomultiplier output and will be used to indicate when

TABLE 1

X	Y	Y CALCULATED	DIFFERENCE	DIFFERENCE (Per cent)
−·1	52	52	0	0
4·8	150	152	+2	+·2
9·8	251	253	+2	+·2
14·8	355	355	0	0
19·9	462	459	−3	−·3
24·8	560	559	−1	−·1
29·9	665	663	−2	−·2
34·8	763	762	−1	−·1
39·9	866	866	0	0
45·0	971	970	−1	−·1
50·2	1072	1076	+4	+·4

Where: X is the measured distance across the record in millimetres accurate to $\pm 0{\cdot}1$mm
Y is the digitization count recorded by the graphic-analogue to digital convertor accurate to ± 1
Y calculated is the value calculated from a straight line fitted to X and Y by the method of least squares

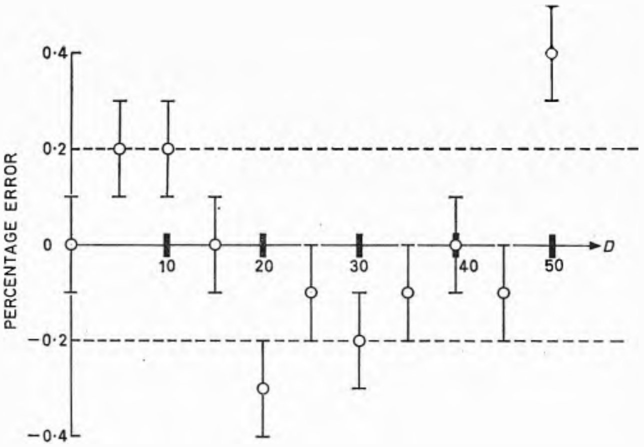


Fig. 4. The percentage error in the digitization as a function of the distance across the record

The dotted lines at the ±0·2 per cent positions indicate the possible error in reading manually the position of the trace and the limits around each point indicate the ±0·1 per cent error due to the ±1 count of the counter-timer

TABLE 2(a)

Results of the digitization of a straight line trace with 3:1 slope (i.e. 72° inclination to the time axis) are shown below.

The column marked READING *indicates the digitization; that marked* DIFFERENCES *indicates the difference between successive digitizations. The variation in the differences is likely due to variation in transport speed of the record. Readings were taken four times per second*

READING	DIFFERENCES
1152	48
1104	47
1057	44
1013	45
968	48
920	48
872	46
826	50
776	45
731	50
681	48
633	51
582	50
532	49
483	50
433	51
382	47
335	45
290	46
244	45
199	48
151	

TABLE 2(b)

The results of digitization of a straight line with 3:1 slope (i.e. 72° inclination to the time axis) are shown below.

The column lists a number of successive digitizations of the same spot on the trace. Readings were taken four times per second

SUCCESSIVE DIGITIZATIONS
644
644
644
644
644
643
643
643
643
644
643
644
644
644
644
643
644
644
644
643
643

Average difference 48
R.M.S. scatter 2·2

there are two traces on the record which means that the record has been scratched. The flip-flop output can signal to the recorder that the particular digitization is questionable.

The system at present has a speed of five conversions per second using a line printer as the digital recorder. Using the device with a high speed recording device and the present 1msec sweep, speeds of almost 1 000 conversions per second would be possible. A speed of several thousand samples per second is possible with a faster sweep and timer, but the real and more practical limitation is the maximum speed at which the record can be transported past the scanning spot.

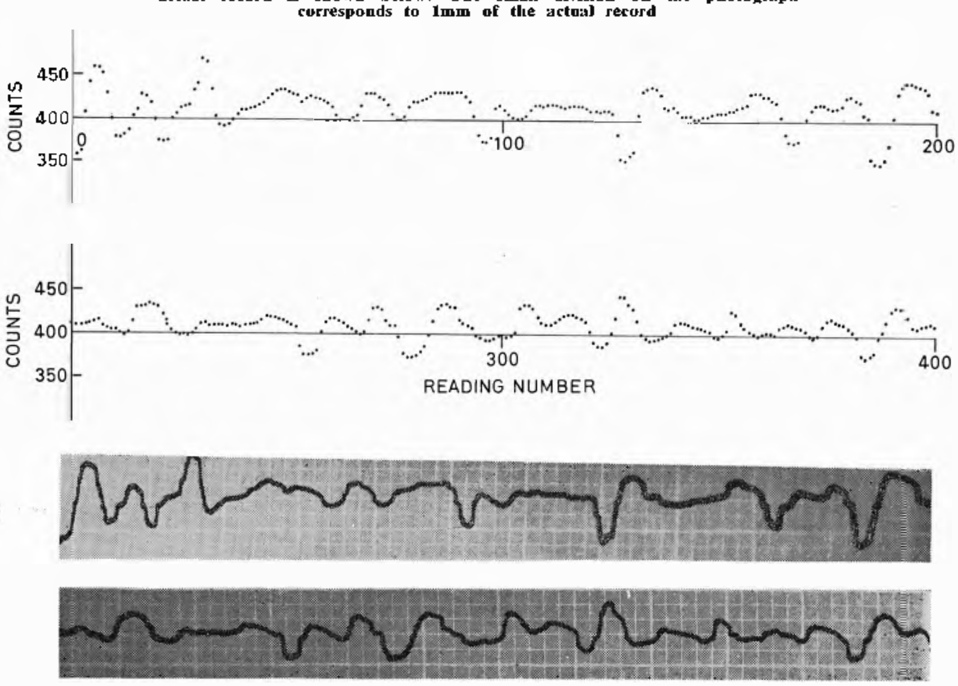
Since only one axis of the c.r.t. is used in scanning the trace, the other could be used to move the spot along the axis of the paper. This would be useful for high speed scanning where paper transport acceleration becomes a problem or for special computation techniques such as averages or autocorrelation.

An attempt will be made to build a scanner which measures the reflected light so that records with a dark trace on a light background can be read automatically.

A graphic-analogue to voltage-analogue convertor is also being built. The spot is made to 'wobble' by superimposing a sine wave on the deflexion plates of the c.r.t. Each time the spot crosses the trace the photomultiplier detects the change in the light transmitted, and by comparison between these pulses and the phase of the sine wave, a voltage can be generated whose amplitude and sign depend on the average position of the spot relative to the position of the trace.

This voltage can be used in a servo feedback loop to move the spot so that its average position corresponds to the position of the trace. Hence, as the strip chart moves the spot will follow the trace, and the voltage on the deflexion plates of the c.r.t. is the voltage analogue of the graphic trace.

Fig. 5. A sample of digitized waveform is shown above. The counter-timer reading is plotted against reading number. An enlarged photograph of the actual record is shown below. One small division on the photograph corresponds to 1mm of the actual record



Conclusions

A graphic-analogue to digital convertor has been built which accurately and rapidly reads strip chart records of the type suitable for Ozalid reproduction. The accuracy is ±0·5 per cent f.s.d. and the repeatability is 0·1 per cent, with a scanning rate variable up to several hundred conversions per second.

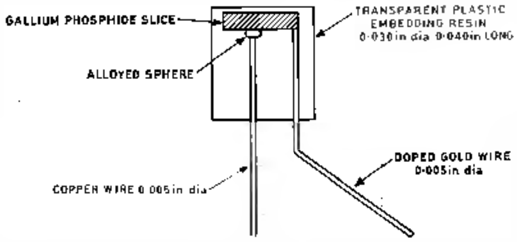
The system of strip chart recorder and graphic-analogue to digital convertor gives the experimenter the benefit of a permanent visual analogue record simultaneously with an easy means of converting the information to digital form.

A Semiconductor Light Source

A new type of light source, based on radiative recombination at a pn junction, and known as a crystal lamp, is shortly to be marketed by the Electronics Department of Ferranti Ltd.

The device, which is believed to be the first of its type produced commercially in the United Kingdom, is in effect a forward biased gallium phosphide diode suitably doped to produce electroluminescent radiation of 7 000Å (red). As can be seen from Fig. 1, the device structure is encapsulated in transparent plastic for the dual purpose of increasing mechanical strength and the amount of light leaving the crystal.

At a forward current of 20mA the red lamp has a voltage drop of approximately $2\frac{1}{2}$V. Under normal d.c. pulsed operation a 1msec pulse at 50mA can be passed through the lamp; however, a 1μsec pulse at 1A can also be passed through it to give twenty times the steady-state brightness.



Construction of the light source

The average brightness is 10 to 40 ft. lamberts, and the lamp is visible under an illumination of 20 ft candles. Switch-on time of the lamp is approximately 0·2μsec and operating temperature range is −20°C to +70°C.

Operating in the reverse biased mode it can be used as a fast light source, light output rise times of less than 3nsec being obtained.

The gallium phosphide red lamp has many potential applications in industry generally, particularly in instrument and indicator applications where its small size typically 0·03in by 0·04in and low operating currents are particularly advantageous. Its most obvious application is as a simple on/off indicating device in transistorized equipment in which the voltage and current levels are suited to it. For example, it could be used to indicate whether part of a complex computer system was on or off, or in a telephone exchange where indication of the state of sub-sections could be extremely helpful for maintenance and fault finding purposes. The numerical output from a digital computer can also be displayed by a linear array of lamps and recorded photographically, while more complex arrays can display more complicated data. Two-dimensional arrays provided with suitable scanning could also perform some of the functions of a cathode-ray tube.

In aerial photography an array of these devices could be used for film marking, and it is of interest to note that a 1msec pulse at 50mA will suitably expose a range of films of various speeds. Used with photoconductive elements, a variety of electronic devices could be constructed. Development of the device was sponsored by Electronics Research Laboratory, Baldock, Hertfordshire, which has been responsible for all the initial research work.

An Integrated Circuit Hearing-aid

Using an integrated circuit manufactured by Texas Instruments, Danavox International A/S of Copenhagen, has recently introduced a new hearing-aid. This is claimed by Texas Instruments to be the first consumer application of integrated circuits in Europe.

The complete electronic circuit is contained in a single wafer of silicon, mounted in the standard flat package, and having the equivalent of six npn planar transistors and fifteen resistors. Only transducers (microphone and earpiece), volume control and battery are required to complete the hearing-aid circuit. The gain is 8dB higher than the previous discrete transistor version, but the major benefits are in size and weight reduction, power consumption and reliability.

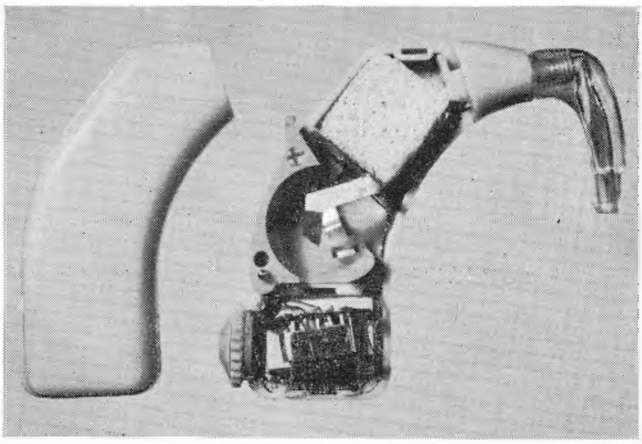
The aid has been measured with a tube length of 35mm mounted directly to the cavity of the 2cm^3 coupler, and with a battery voltage of 1·5V.

Tube dimensions: Inner diameter: 2·0mm
Outer diameter: 3·2mm

Typical data is as follows:

Maximum acoustic amplification: 58dB
Maximum acoustic amplification at 1 000c/s: 45dB
Maximum acoustic output: 124dB re 0·0002μbar.
Maximum acoustic output at 1 000c/s:
124dB re 0·0002μbar
Random noise at gain 40dB at 1 000c/s:
Less than 65dB re 0·0002μbar
Rated acoustic output (10 per cent distortion) at 1 000c/s:
123dB re 0·0002μbar.
Earphone/Receiver: Magnetic internal
Special super powered integrated circuit.
Volume control: Separate continuous, attenuation range more than 40dB

The integrated circuit hearing-aid



A Track-Selection System for a Magnetic Drum Store

By S. C. Acharyya*, Ph.D., B.E.E.

This article deals with the design of a track-selection switch for use with a magnetic drum storage system. The use of transistor gates in the switch imposes some restrictions upon the selection of the component values, especially when two voltage levels are used.

With low-inductance head windings, the switch arrangement proves to be simple, reliable and economical. With a magnetic drum having high head winding inductance, the necessity to use high voltage transistors makes the switch comparatively more expensive, although the simplicity and reliability of the arrangement are retained.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

THE purpose of using a track-selection switch is to allow information in the form of coded pulses to be routed to or from a selected recording head out of a number of heads of the magnetic drum. The arrangement allows only one read and one write amplifier to be used for recording information, at any one time, to or from any of the tracks included in the matrix.

The track-selection switch, as discussed in this article, uses electronic switching with the associated advantages of rapid access track-selection with reliability. Attempts have been made to combine the advantages of various other available track-selection methods[1,2]. In the present selection switch, a desired track is selected by direct connexions to the appropriate lines of row and column selecting decoder outputs, thus making an economy in the number of components used.

The performance of the switch has been tested with a 4 × 4 matrix formed out of 16 heads of a magnetic drum whose head windings were originally designed for use with valve circuits. The use of valve circuits allows the head winding inductance to be on the high side since the problem of raising appropriate write current in the windings can more easily be solved by use of suitable transformer and voltage pulses. Special problems have to be faced when transistor circuits are used with such head windings. The various problems arising in the design of the switch are discussed in the following sections.

Circuit Description

The circuit diagram of the switch is shown in Fig. 1. The transistors as shown in the diagram are either saturated or cut off, depending on whether the associated selection line is at $-V_{cc}$ or at $+V_B$ volts. This type of operation ensures that the transition time of the switch depends on the cut-off frequency and storage time of the transistors.

The voltage shown for the selection lines corresponds with the selection of the track associated with recording head H_{00}. *X*-select transistor VT_{x0} is saturated, raising the potential of the centre-taps of all the head-transformers in the selected row to approximately earth potential. *X*-select transistors of other rows are cut off by $+V_B$ volts at their bases. The *Y*-select transistor VT_{y0} is cut off, allowing transistors (VT_G) of the transformer associated with head (H_{00}) to conduct. The bases of the other *Y*-select transistors are held negative so as to saturate them. The *Y*-select lines of the unselected columns are held at $+V_B$ volts which bias the transistors (VT_G) of the unselected columns to cut-off.

For reading, the write amplifiers, e.g. two npn transistors (VT_3), are held cut off and a read signal raises the potential at the base of npn transistor (VT_4) allowing it to conduct. A low impedance conducting path is thus established through transistors (VT_G) of selected head (H_{00}) and transistor (VT_4) so as to allow voltage signals from the selected track to appear at the primary of the read transformer, the secondary output of which is then applied to the read amplifier.

During writing, a write command normally opens a gate (not shown) which allows write and inverse write waveforms (in phase modulation method of writing) to operate the two npn transistors (VT_3) allowing appropriate write current signals to pass through the two halves of the primary windings of the selected head-transformer. By transformer action, these current signals are transferred to the selected head winding which records the information on to its associated track.

Diodes (MR_1) prevent transistor (VT_4) from being overstressed by difference voltage ($V_{cc1} - V_{cc}$) between its emitter and collector during writing operation when the two collector lines of transistors vary (approximately) between earth and $-V_{cc1}$ potentials.

Design Considerations

Details of the recording heads of the magnetic drum used in the test are given in Fig. 2. Each of the head windings has been connected to the track-selection circuit by a head transformer. The centre-tapped primary of the transformer allows a balanced configuration of the circuit to be worked out. A balanced configuration helps cancellation of extraneous noise and switching transients and simplifies the design of the write circuit for the phase modulation method of writing digits. It also permits the use of direct current flow through the transformer windings without saturating the cores.

SYMBOLS

I_w	=	Write current flowing through the primary half winding during writing.
$-V_{cc}$	=	Negative voltage supply to all transistors except write amplifiers.
$-V_{cc1}$	=	Supply voltage for npn transistors in write-circuit.
V_{ce}	=	Transistor collector to emitter voltage drop.
α_0'	=	Common emitter current gain (d.c.) for the transistor.

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D.C. Design

In order that the arrangement shown in Fig. 1 can work as a track-selection switch with reliability, the component values have to be selected so that correct voltage levels are maintained in the circuit at any one time. The considerations that lead to the establishment of such voltage levels are given below.

Value of resistance (R_1) at the base of X-select transistors should be such that:

$$\frac{V_{cc}}{R_1 + R} \times \alpha_{0(\min)}' > I_W \quad \ldots\ldots\ldots\ldots \quad (1)$$

where

$$I_W \simeq V_{cc1}/R_3 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

This prevents X-select transistor (VT_{x0}) of the selected row from coming out of saturation when heavy write current flows through it.

R_2 is selected from two considerations. Firstly, R_2 has to satisfy the condition that:

$$\frac{V_{cc}}{2((R_2/2) + R_4)} \times \alpha_{0(\min)}' > I_W \quad \ldots\ldots\ldots\ldots \quad (3)$$

Secondly, R_2 is dependent on the value of other resistances (R_3, R_4 and R_5). This dependence follows from the requirements that the potentials at different points of the circuit have to be maintained at certain values during reading the information from the selected track. During reading, the npn transistor (VT_4) is bottomed, and a base current of V_{cc}/R_6 flows through it. A current I_R flows through the selected line where:

$$I_R = \frac{V_{cc} - (V_{ce}\text{ of }VT_{x0} + V_{ce}\text{ of }VT_G + \text{voltage drops across diodes }MR_1 + V_{ce}\text{ of }VT_4)}{R_5 + (R_3/2)} \quad \ldots \quad (4)$$

This current flow causes the common read/write line to be at a potential V_{cR} where:

$$V_{cR} = -(V_{ce}\text{ of }VT_{x0} + V_{ce}\text{ of }VT_G + (I_R/2) \times R_3)\text{ volts} \quad \ldots\ldots\ldots\ldots \quad (5)$$

The potential of the selected column line (Y_0 in the case shown in Fig. 1) is approximately

$$V_Y = -\left(V_{ce}\text{ of }VT_{x0} + V_{bc}\text{ of }VT_G + (R_2/2) \times \frac{V_{cc}}{R_4 + (R_2/2)}\right) \quad \ldots\ldots\ldots\ldots \quad (6)$$

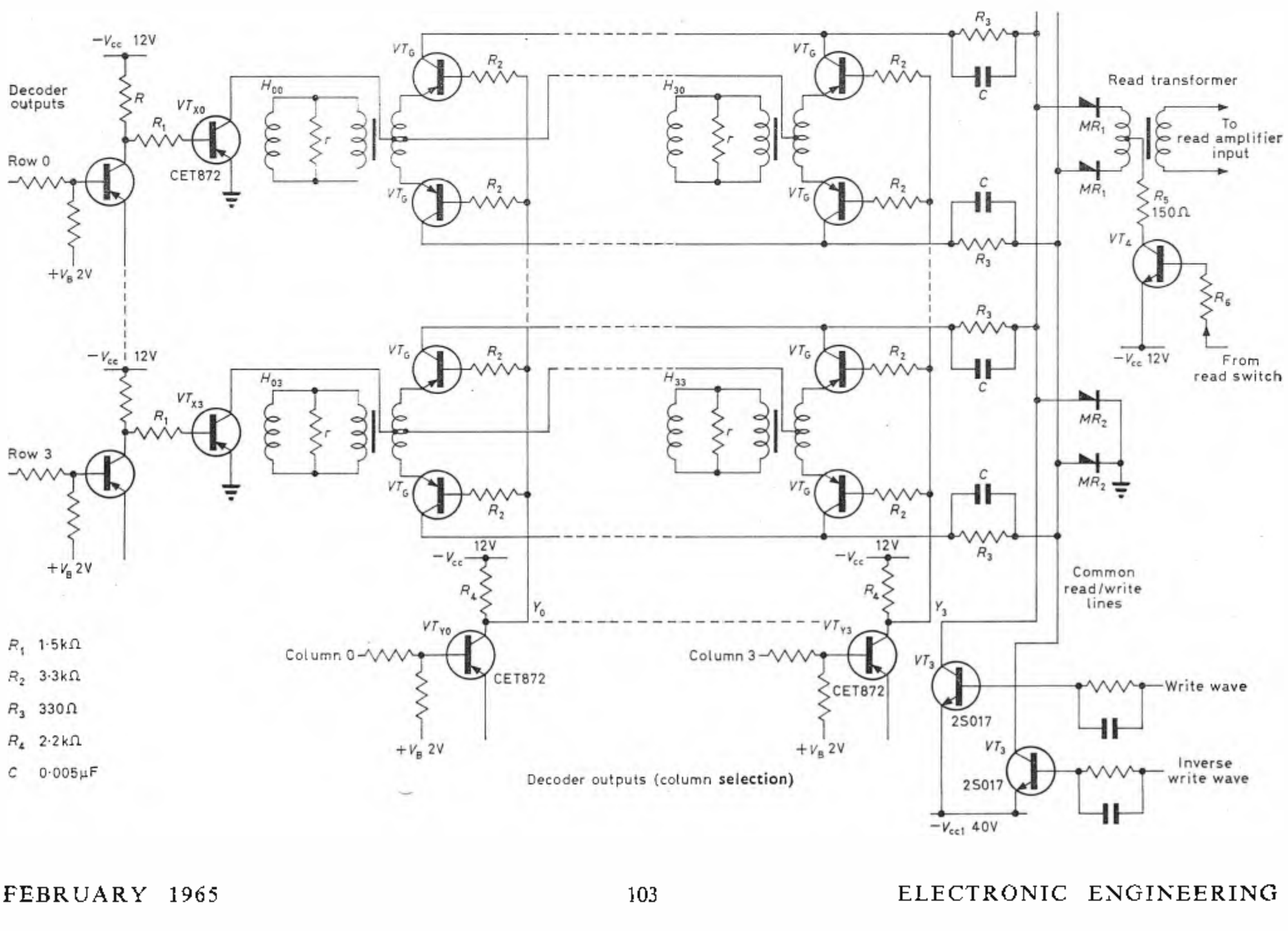
For the matrix to work satisfactorily, V_{cR} and V_Y have to be such that V_{cR} is at a more negative potential than that of V_Y, i.e. $V_Y > V_{cR}$. Otherwise, collector to base potentials of transistors (VT_G) of the selected column and unselected rows become forward biased and cannot offer a high impedance for blocking the unselected signals from passing into the common read lines. The crosstalk between selected and unselected heads becomes objectionable. The circuit should be designed so that the difference in magnitude between V_{cR} and V_Y is a minimum of 500mV; the greater the difference the better the noise rejection.

Since the value of R_3 is selected to fulfil the requirements of a heavy write current, its value is generally low. Resistance (R_5) is therefore used to limit the current flowing in the selected circuit during reading. Too heavy a current flowing through the transformer windings tends to deteriorate the signals from the magnetic heads at the output of the read transformer.

Head Transformer

A head transformer has been used to:

Fig. 1. Circuit diagram of track-selection system



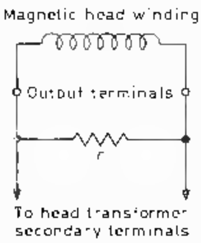
(1) Step up, during reading, the output signal from the head windings before being applied to the read transformer primary.

(2) Step up the current in the head windings during writing so that the current ratings of the transistors on the primary side of the head-transformers may be kept down.

(3) Obtain a balanced configuration for the circuit by the use of a centre-tapped primary of the transformer.

The design of the transformer appears to be slightly difficult when due consideration has to be paid, firstly, to the availability of transistors of appropriate current and voltage ratings within reasonable cost; secondly, to the effect of inductance of the head windings after transformer action upon the current rise time during writing operations and, thirdly, to the effect of regulation of the transformer and head windings upon the magnitude of read signal voltage.

A high value of transformation ratio n tends to increase head inductance by n^2 when transferred to the primary side. Since the rate of current rise in an inductive circuit is given by $E = L(di/dt)$, any increase in n tends to increase the voltage ratings of the transistors (VT_3) by an amount which is proportional to n^2 for a required rate of current rise. On the other hand a low turns ratio tends to decrease the read signal output from the transformer and also requires heavy current ratings for the npn transistors (VT_3).



Recording head winding—a single coil with two output terminals
Measured inductance of the coil ≃ 300μH
Write current = 150mA change (peak-to-peak) (by experiment)
Open circuit read signal (minimum value) = 8mV (peak-to-peak)
Digit period ≃ 6·2μsec
Available time for 150mA current change ≃ 3·1μsec
Damping resistance (r) = 1kΩ

Fig. 2. Recording head characteristics and data relevant to head transformer design considerations

The actual number of turns in the transformer windings and the particular type of core are determined from considerations of inductance of the transformer windings and its regulation, and also the effect of any unbalance in the centre-tapped winding upon saturation of the core.

After experiments with various combinations of types of ferrite cores, number and ratio of turns and different transistors, a compromise design has been arrived at for the head transformer, particulars of which are given in Fig. 3.

Performance

Current Rise

In order to record information at a p.r.f. of about 160kc/s in a phase modulation system the available time for change of current direction (peak-to-peak) in the head windings is only half the digit period, i.e. 3μsec. In order to obtain an open-circuit read signal (minimum) of about 8mV (peak-to-peak) a current change of about 150mA in 3μsec is required in the head windings. Due to the effects of inductances of various windings and the time-constant on the primary side, a current change of about 120mA only is obtained in 3μsec by application of −40V. By putting suitable capacitors across the current limiting resistors (R_3) a better current rise time has been achieved. The use of the capacitors across the resistors allows the peak voltage across the transformer winding to be maintained for a longer period so that the current rise time approaches a sawtooth rather than exponential wave shape. This allows a higher magnitude current to be attained in half the digit period. With −40V supply, a current rise over 150mA (peak-to-peak) has been obtained in the windings of the magnetic heads of the drum.

Crosstalk

The crosstalk during reading among the selected and unselected heads is minimized by applying a reverse bias at the bases of the transistors (VT_G) associated with unselected head transformers. This can be achieved by designing the X and Y select decoder output transistors with their emitters connected to $+V_B$ supply. Also attention has to be paid in the d.c. design of the circuit (as has been discussed in a previous section) to minimize the effect of crosstalk.

Switching Time

The speed of track selection is limited by the fact that the transistors are heavily saturated so that storage time is appreciable. With GET113 transistors, the transition time has been measured at 11μsec.

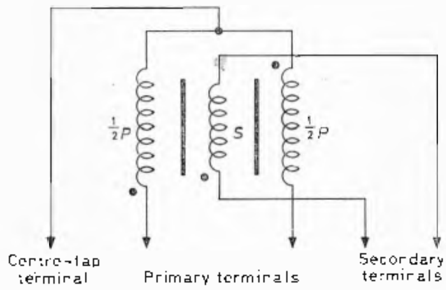


Fig. 3. Head transformer winding details

½ primary: 45 turns; secondary: 30 turns
36 s.w.g. Lumex wire used for the windings
Type of core: FX1105 or FX1239 (E type)

To determine the overall track-selection speed, the recovery time of the associated read amplifier has to be taken into account. The switching transient exhibited by the switch consists of a short spike of about 1V magnitude and 3μsec duration when measured at the secondary winding of the read transformer. The read amplifier used takes 10μsec to come out of saturation due to the voltage spike appearing during the switching transient. The overall switching time during reading is, therefore, about 21μsec which is the recovery time of the amplifier plus the transition time of the transistors in the switch.

Remarks on Test Conditions

The performance of the track selection system has been tested with different component values and with different types of transistors. During reading, the system using GET113 or GET 872 gives satisfactory results for a 4 × 4 matrix. During writing, the transistor ratings require to be changed to different values because of the low collector to base voltage rating of these transistors. While writing, the common lines of the matrix vary between $-V_{ee1}$ and near earth potentials. Using above 40V for V_{ee1}, the transistors (VT_G) of the non-selected head transformers are subjected to overvoltage while writing. This causes spurious writing on the unselected tracks and generates noise while reading.

When voltages in the range of 40V or above are used for V_{ee1} because of special requirements for writing, more expensive transistors (OC23) used in the switch prove to be very satisfactory during both reading and writing.

However with a lower inductance head winding having a centre-tap available, circuits can be designed using lower voltage for V_{cc1} and using inexpensive transistors.

Conclusion

The switch has not undergone any rigorous life tests, but it is expected that, with good quality transistors, its performance should improve. With the expansion of the matrix to include a greater number of tracks, the quality of performance may tend to decrease slightly due to the increased effects of crosstalk from additional tracks but, with proper design and selection of components, satisfactory performance can be expected with a matrix including more than 100 tracks. Also, a much better signal-to-noise ratio can be achieved by using a magnetic drum having a magnetic surface of superior characteristics than the one that has been used.

Acknowledgment

The author is indebted to those who were of help to him during the development of the track-selection system at the Power Systems Laboratory of the Manchester College of Science and Technology.

REFERENCES

1. BOOTH, A. D. High Speed Track Selection for a Magnetic Drum Store. *Electronic Engng.* 32, 209 (1960).
2. SEADER, L. D. Magnetic Recording Head 'Selection Switch'. *I.B.M. J. Res. Developm.* 2, 36 (Jan. 1958).

An Educational Computer

This educational aid has been produced by Elliott-Automation Ltd in conjunction with the Electrical Engineering Department of Battersea College of Advanced Technology to demonstrate the way in which a digital computer performs routing, control and arithmetic operations on binary numbers. Input and output of binary quantities is by push-buttons and lamps, while selection of the function defining the operation mode to be performed on these quantities is by a multi-position switch.

This unit can be considered as the basic part of the central processor of a digital computer. Additions to the logic can be made in order that further features of the digital computer, can be added such as:

(1) Core store or drum store.
(2) Control registers and function mesh (to replace the function switch and introduce simple programming facilities).
(3) Tape reader and input logic to store.
(4) Tape punch and output logic from store.

A more sophisticated and gradually improved educational aid can be produced in this manner.

The prominent feature of the design of the basic computer unit is that all logic configurations are readily accessible on mounting board cards of three to six Minilog elements, to enable the student computer logic designer to investigate each transistorized circuit. At the same time, any modifications and improvements to the logical operation of this unit can be speedily implemented by the addition of Minilog elements or extra boards of elements. Thus, planned development projects can be incorporated on the same unit. This basic computer can be achieved by either the purchase of a kit of Minilog elements and logic circuits or the purchase of a complete kit of Minilog elements, boards, chassis, cabinet, power supply units, display panel and the logic circuits, or finally the purchase of a completely tested and commissioned unit.

The unit is a synchronous type with a clock frequency of 5kc/s for normal operation, 1c/s for demonstration, and a further facility of '1 shot,' i.e. one clockpulse per initiation of a push-button.

The operation is in serial binary form with a word length of 16 bits. Associated with each word time are 20 clockpulses, 16 of which are to operate the arithmetic logic and four clockpulses named 'gap times' which are included to enable program or data to be transferred from the store (when included) to the computer.

Input of data in binary code to the store buffer register and accumulator is by push-buttons, while output of the contents of these two registers is by lamps. There are seven functions available with the basic unit, but there is, of course, complete flexibility for additions to suit the user's individual requirements. There are three arithmetic registers included, and these are as follows:

(1) Store buffer register (A) (16 bits)
(2) Accumulator register (B) (16 bits)
(3) Test register (C) (16 bits)

The seven functions which make use of the above registers are as follows:

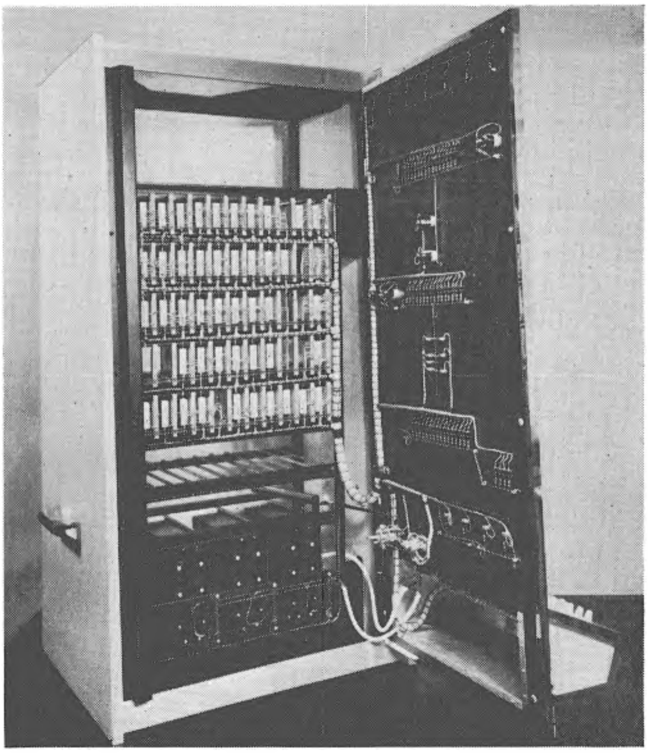
(a) *Load:* Transfer A to B clearing A.
(b) *Add:* Add A to B clearing A.
(c) *Subtract:* Subtract A from B clearing A.
(d) *Shift:* $B \times 2^n$ where n is the number in the shift control register.
(e) *Negate:* Complement A transfer to B clearing A.
(f) *Multiply:* $B \times A$ to B clearing A (C holds partial product during multiplication).
(g) *Transfer:* B to A clearing B.

To enable these functions to be ordered, a switch is used in conjunction with an 'OBEY' switch which causes the order to be obeyed.

Should the user wish to construct configurations in logic which differ from those of the basic machine, this can be done by simply inter-changing connexions and plug-in boards.

It is expected that students will first be given a theoretical appraisal of how computers work which can then be demonstrated on the machine. The instructor can insert faults for the student to correct. He can remove entire logic sections, leaving the student to reconstruct the machine from a plug-in board, suitable components and a performance specification. In this way the student will have been given experience of all the basic problems possible with a digital computer, and a real understanding of computer logic, organization and design.

The interior of the educational computer



Improved Thermal Delay by New Circuit Principle

By J. E. Brooke-Stewart* and G. F. Klepp*

Thermal delay switches find applications in the delay of the connexion of h.t. to indirectly heated valves and rectifiers, in motor starting, repetitive switching, and numerous other delay circuits.

Thermal delay switches have two main characteristics, the making time and opening time. If the final delay is achieved not by the making time alone, but by the addition of the making time and the opening time, then the timing spread between individual switches is reduced as compared with either of these two characteristics alone. The combined delay time is shown to be largely independent of heater voltage, ambient temperature, and to have less variation with switch life.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

THE thermal delay switch was developed as a device for switching on equipment after a specified time from closing a main switch. Delay is usually achieved by the well established bi-metal action. The operation of a typical switch is explained by reference to Fig. 1.

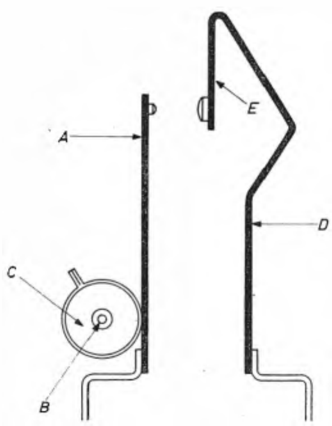


Fig. 1. Action of thermal delay switch

A bi-metallic strip *A*, the active strip, deflects under the influence of a wound wire heater, *B*, enclosed in a ceramic tube *C*, and makes contact with a second bi-metallic strip *D*, the stationary strip. Following closure, any heat transference from the active strip into the stationary strip causes the reversed portion, *E*, of the stationary strip to deflect towards the active strip. Contact pressure is thereby increased. Any variation in ambient temperature causes both strips to deflect in the same direction, thereby maintaining the gap relatively constant. Complete compensation over a wide temperature range is usually difficult to achieve with an otherwise acceptable configuration of two bi-metallic strips. The simplest totally compensated arrangement of two bi-metallic strips which comes to mind is that of two identical strips, side by side. Practical conditions often preclude the application of an ideal arrangement. For the switch now under consideration, the compensation is effective only over the limited range shown in Fig. 11.

Two parameters may be ascribed to the thermal delay switch, the making time and the opening time. The making time of a thermal delay switch is defined as the time from the application of the heater voltage until the instant at which the contacts make. The opening time of a thermal delay switch is defined as the time measured from disconnexion of the heater supply until the instant of contact opening. The opening time is dependent upon the time of heater voltage application, but will reach a steady value when thermal equilibrium has been established within the switch envelope.

Thermal delay switches can cover a range of delay times. Table 1 is an extract from a thermal delay switch data sheet. These switches are sealed into valve type envelopes with inert gas filling.

TABLE 1

SWITCH	HEATER VOLTAGE (*V*)	HEATER CURRENT (*A*)	MINIMUM DELAY (*sec*)	MAXIMUM DELAY (*sec*)	CONTACT RATING ON OPENING
S102/1K	6·3	0·5	44	66	100mA at 50V
S102/2K	6·3	0·5	44	66	1A at 250V
S103/1K	27·0	0·115	36	54	100mA at 50V
S104/1K	6·3	0·5	25	35	100mA at 50V
S104/2K	6·3	0·5	25	35	1A at 250V
S105/1K	27·0	0·115	20	30	100mA at 50V
S106/1K	19·0	0·165	44	66	100mA at 50V
S107/1K	6·3	0·5	8	15	100mA at 50V
S108/1K	24·0	0·130	36	44	100mA at 50V

The nominal range may be extended by a suitable choice of heater voltage. Fig. 2 shows the making time characteristic related to heater voltage of a typical thermal delay switch, the S102/1K. Fig. 3 shows the relationship between opening time and previously applied heater voltage for the S102/1K.

For still longer delay times, a number of switches may be operated sequentially, the final delay being the sum of the individual making times. The addition of several making times, however, also entails the addition of individual deviations from switch to switch. The addition of three S102/1K, for example, would produce a delay of 165 ± 33sec.

A study of quality control records for the production of thermal delay switches showed that switches with high making times had low opening times. The sum of making and opening times proved to be relatively constant. Fig. 4 shows the distribution of making time, opening time, and the sum of these two, plotted on a probability scale. Between 5 per cent and 95 per cent of the product, the percentage spread of these three values is 21, 53 and 13 per cent respectively.

Figs. 2 and 3 clearly show the reciprocal nature of the making and opening times.

The effect may be explained by considering the rate of deflexion of the active strip. Measurements of deflexion have been made on a thermal delay switch with the

* *Standard Telephones & Cables Ltd.*

stationary strip removed. These measurements are plotted as Fig. 5. It can be seen that the strip takes a fundamental time to reach the steady state, and that this time is practically independent of input energy. It is only the rate of change of deflexion, and total deflexion which vary with input. The strip under consideration requires 204sec to reach this state of equilibrium. If the contact gap be taken as 0·035in, it can be seen that with 5·5V applied to the heater, the making time is 75sec. For the remainder of the 204sec, the strip will build up a contact pressure proportional to the restrained deflexion A. When the heater supply is removed, the strips will not open until the restrained force has become zero. If the making time is reduced to 34sec by increasing the heater voltage to 7·0V, the restrained force will increase by an amount proportional to the restrained deflexion C. The opening time will be proportionally longer. Thus, if the making time is reduced, the restrained force will increase and the opening time become longer.

Circuit for the Addition of Making and Opening Times

The constancy of the sum of making and opening times can be applied to gain a very constant delay. Fig. 6 is given as a typical circuit to demonstrate the principles involved. If the making times of the three switches are labelled M_1, M_2 and M_3 respectively, and the opening times O_1, O_2 and O_3 respectively, then the circuit will add these times in the sequence $M_1+M_2+M_3+O_1+O_2+O_3$. Fig. 7 shows the effect of adding the making times and opening times of three S102/1K switches operated over a wide range of heater voltage. The addition of five S102/1K making times is included as a comparison. The striking result is the manner in which a thermally operated device,

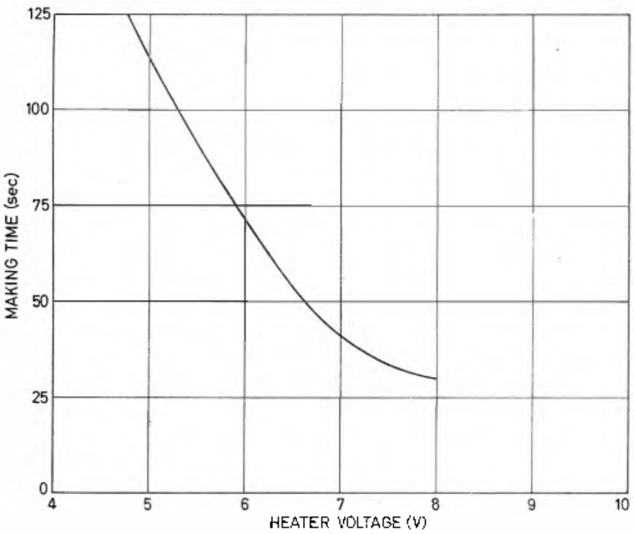
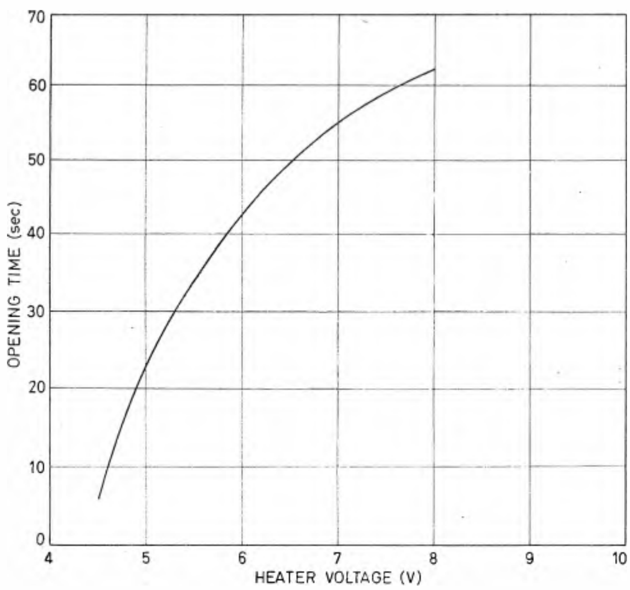


Fig. 2. Making time-heater voltage for S102/1K

Fig. 3. Opening time-heater voltage for S102/1K



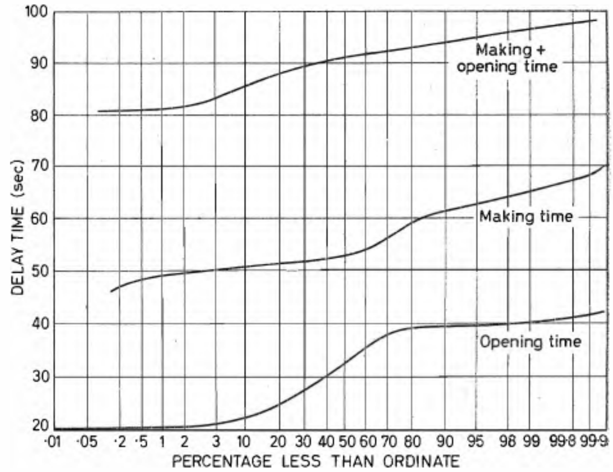


Fig. 4. Spread of S102/1K

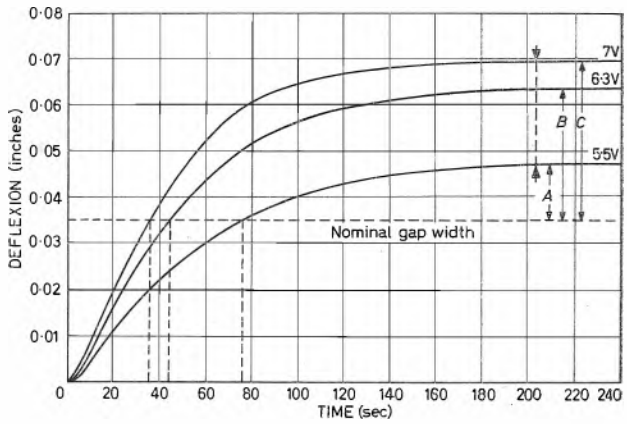


Fig. 5. Deflexion time characteristic

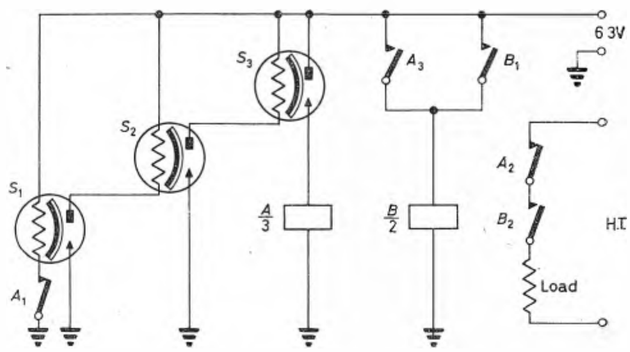


Fig. 6. Typical circuit for the addition of making and opening times

Sequence: S_1 makes and switches on S_2; S_2 makes and switches on S_3; S_3 makes and connects relay A; A_1 opens and switches off S_1; A_2 opens; A_3 closes and connects relay B; B_1 closes; B_2 closes; S_1 opens and switches off S_2; S_2 opens and switches off S_3; S_3 opens and switches off relay A; A_2 closes and connects the load which might be expected to be critically dependent upon input energy, is shown to have a practically flat characteristic over a wide range of heater voltage. The following analysis shows that the degree of flatness of the characteristics is related to the rate of change of the temperature of the bi-metallic strip during the heating and cooling periods.

Delay = Making time + Opening time

i.e. $D = M + O$

Consider a change in temperature of the bi-metallic strip due to a change in heater voltage.

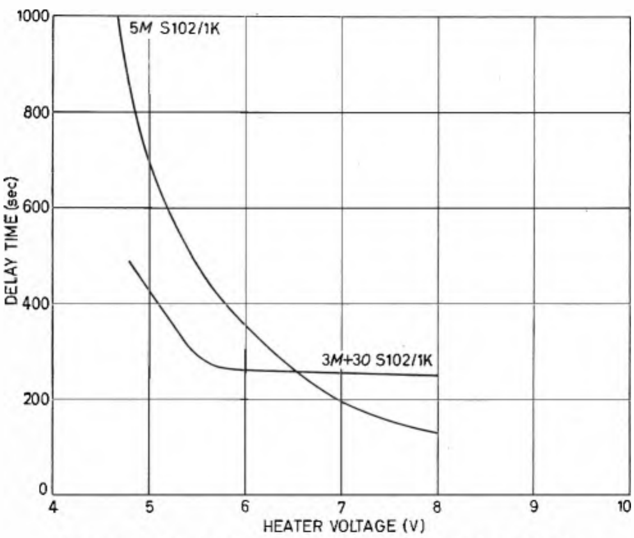


Fig. 7. Addition of making and opening times, S102/1K

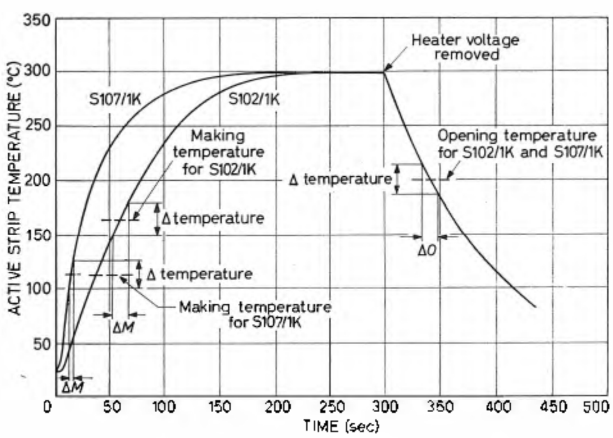


Fig. 8. Active strip temperature-time (heating and cooling)

Change in Delay = Change in Making time − Change in Opening time.

i.e. $\Delta D = \Delta M - \Delta O$ since the making time and opening times are inversely proportional.

If $\Delta M = \Delta O$ then $\Delta D = 0$, i.e. the characteristic is flat.

If $\Delta M < \Delta O$ then ΔD is negative, i.e. a decrease in heater voltage causes a decrease in delay.

If $\Delta M > \Delta O$ then ΔD is positive, i.e. a decrease in heater voltage causes an increase in delay.

The change in making time is inversely proportional to the rate of change of the active strip temperature. This rate of change can be varied by increasing or decreasing the thermal conductivity of the ceramic tube C in Fig. 1,

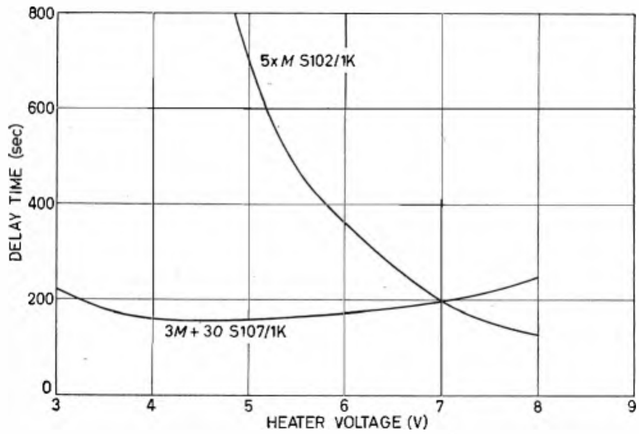


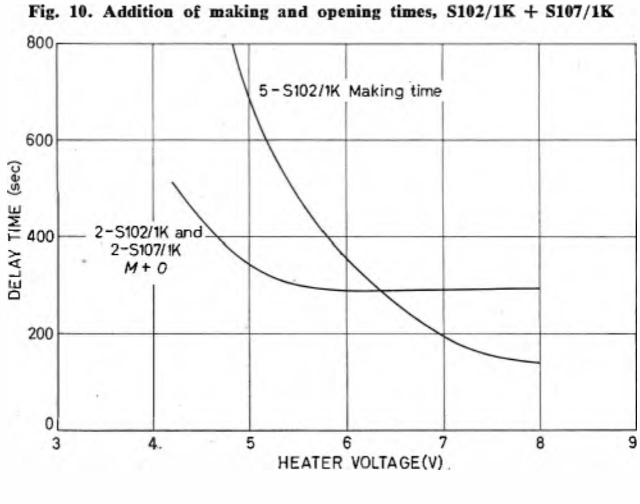
Fig. 9. Addition of making and opening times, S107/1K

e.g. by altering its thickness. The change in opening time is dependent upon the rate of loss of heat from the active strip, and is substantially unaffected by ceramic thickness, i.e. the rate of heating can be varied without affecting the rate of cooling. There is an optimum thickness of ceramic which will produce a rate of change of temperature on heating equal to the rate of change of temperature on cooling. Fig. 8 shows the variation in active strip temperature during the heating and cooling periods for two thermal delay switches, the S102/1K and the S107/1K. These switches are identical with the exception of the thickness of the ceramic tube, this being 0·066in for S102/1K and 0·025in for the S107/1K. It is seen that in the case of the S102/1K, ΔM is greater than ΔO producing the positive change in total delay shown in Fig. 7. With the S107/1K, ΔM is less than ΔO, the condition for a negative change in total delay. Three S107/1K switches were operated in the circuit of Fig. 6. The resultant curve is given as Fig. 9, where the negative characteristic is confirmed, and the independence of heater voltage even more striking than in Fig. 7. Fig. 10 shows the result of combining two S102/1K and two S107/1K. This produces a very flat characteristic between 6 and 8V.

Variation of Delay Times with Ambient Temperature

It was mentioned in the introduction, that the ambient temperature compensation may not be perfect owing to the physical layout of the two bi-metallic strips. As a consequence, the timing of a thermal delay switch can have a temperature coefficient varying in sign with ambient temperature. A typical thermal delay switch, the

Fig. 10. Addition of making and opening times, S102/1K + S107/1K



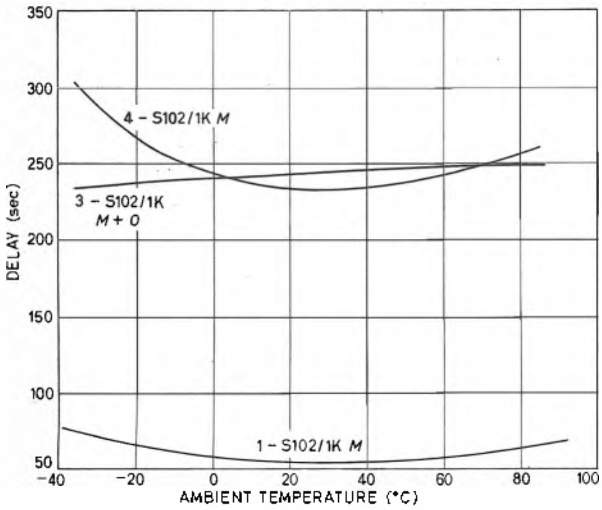


Fig. 11. Ambient temperature compensation with addition of making and opening times

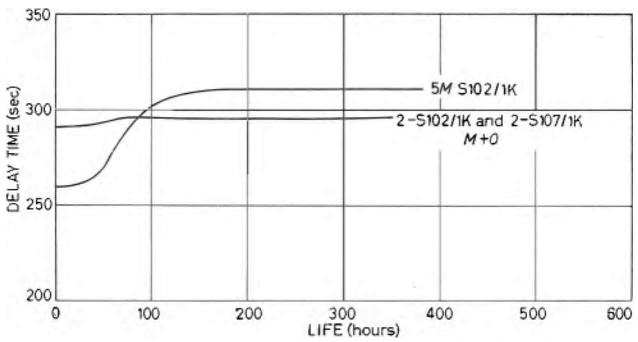


Fig. 12. Variation of delay time with life

S102/1K, may have a making time of 58sec at +23°C, 75sec at −35°C and 65sec at +85°C. If the addition of making and opening times is considered over the same range of temperature, the total delay time is observed to be less dependent upon ambient temperature than are the two individual times. Fig. 10 shows that a delay time of 230sec, achieved by adding the making times of four S102/1K switches, will have a spread of 80sec, i.e. 35 per cent. The same delay time achieved by the addition of the making and opening times of three S102/1K will have a spread of only 14sec, i.e. 6 per cent. This is a significant improvement in the ambient temperature behaviour of thermal delay switches.

Change of Delay Time with Life

A thermal delay switch may show small changes in the value of the making time during early operational life. The change is due to a redistribution of stresses, unavoidably introduced into the bi-metal during the manufacture of the switch. For example, the S102/1K can have an increase in making time of up to 10sec during early life. The increase in making time is accompanied by an decrease in the opening time and compensation by addition is again possible. Fig. 11 shows the variation with life, of delay time achieved by the addition of the making times of five S102/1K switches. The figure also shows the very much reduced variation with life in delay time achieved by the addition of the making and opening times of two S102//1K and two S107/1K switches.

Conclusion

It has been shown that by the use of a novel circuit, in which thermal delay is achieved by the adddition of making and opening times, it is possible to achieve delay times of several minutes, with an appreciably reduced product spread, improved ambient temperature compensation, improved life performance and an independence of heater voltage over a wide range as compared with conventional methods. The principle of operation is not confined to long delay times. Two S107/1K switches, with nominal making times of 11·5sec each, will give more constant delay times of the order of 40sec with this mode of operation, than, for example, the S103/1K with the nominal making time of 45sec.

The next step towards new design arising out of the above work was the consideration of multiple switches in a single unit. There is now a unit commercially available, coded SS110/1D, which consists of two mechanically separate switches sealed in a common glass envelope. When used in a circuit which gives a delay comprising the sum of the two making times and the two opening times, the switch will give a delay of about 240sec with good heater voltage and ambient temperature independence.

Eye Movement Recording

A transistorized data processing system specially developed by The M.E.L. Equipment Co Ltd for Reading University is being used in the study of visual perception.

The human eye, in addition to being capable of voluntary movement, makes a number of small involuntary movements when 'fixating,' that is when the gaze is fixed on a stationary point. These involuntary movements are of three classes: drift, saccades and tremor, and their magnitudes are subject to individual variation. The drift is a slow movement at a velocity of about 5 min arc/sec. Saccades are fast movements which tend to compensate for drift and are typically of 40msec duration, 6min arc median magnitude. Tremor comprises small amplitude oscillations of 20sec arc with a frequency range of approximately 20 to 150c/s. The correlation between these various movements is at present being studied in the Physics Department at the University.

The person on whose eye the measurements are to be made steadies his head by resting it against a frame and biting on a fixed dental mould. Also, he wears a contact lens to which is attached a plastic stalk carrying a small mirror. Light is reflected from the mirror via an optical system on to two photomultiplier tubes. The system is designed so that the amount of light falling on one photomultiplier is proportional to the vertical component of the eye movement, while the light on the other photomultiplier is proportional to the horizontal component.

The outputs from the two photomultipliers are recorded on separate tracks by a high speed magnetic tape recorder at speeds of up to 60in/sec. The data is subsequently converted into binary coded decimal form on punched paper tape so that it may be presented to a computer for analysis.

The data conversion is carried out by the M.E.L. equipment. First, one track is completely processed and then the other. The tape is replayed at a minimum speed of $1\frac{7}{8}$in/sec which gives a maximum reduction of 1:32 of the recording speed. The signal is passed to a d.c. amplifier, and very short samples of 0·8msec duration are taken, the sampling rate being determined by a clock signal previously recorded on the tape. The samples are digitized and passed to a register, and the flow of data is controlled so that it does not exceed 110 characters/sec, which is the maximum speed of the punch.

In addition to the tape read-out, the equipment incorporates a digital printer for use when sampling manually, and a neon tube binary coded decimal display for checking and setting up the overall system.

The punched tape provides a permanent record of the measurements from which analytical information, such as the median saccadic movement of the eye, can be readily obtained by an appropriately programmed computer.

A Complementary-Pair Cascode for H.F. Amplification

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

It is generally accepted that the grounded emitter/grounded base cascode circuit can be used as an alternative to neutralization. This article compares the advantages of cascode and neutralization, and proposes a complementary-pair cascode circuit which is shown to have advantages over the normal cascode circuit, especially at frequencies above 10 to 20Mc/s.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

ONE of the disadvantages of transistors when used as amplifiers is that they are not unilateral devices: that is to say they have internal feedback from output to input. The effect of this is twofold. Firstly, any change in the load impedance connected to a transistor changes the input impedance (and vice-versa, any change in the source impedance changes the output impedance). Secondly, the internal feedback, which is negative, reduces the maximum available gain which may be obtained from the stage. Because the feedback is reactive, a point will be reached at high frequencies at which the stage will oscillate if the (forward) stage gain is made too high. At low frequencies the feedback, being capacitive, has a very high impedance and can normally be neglected.

Unless something is done about the internal feedback when operating at higher frequencies, the available gain will be reduced, and any attempt to increase the gain will tend to produce instability. Furthermore, the tuning and alignment of the amplifier will be much more difficult because the process of tuning changes the impedance of the tuned circuit. If the tuned circuit forms the load impedance of a transistor, the input impedance will change as the load is tuned. This input impedance however is connected across (or across part of) the previous tuned circuit which will thereby also be detuned. The effect is cumulative, and it becomes very difficult to achieve the required overall pass-band.

Neutralization

It is possible to eliminate most of the detuning effect by mismatching[1]; that is by adding sufficient resistance and capacitance in parallel with the tuned circuit to make any change of input impedance negligible. This method however is rather wasteful of gain.

The usual method of overcoming the effects of internal feedback is to unilateralize[2,3] or to neutralize the transistor stage. Basically this involves feeding back externally from output to input a current of equal amplitude but opposite phase to that fed back internally. Unilateralization is an exact process in which the compensation is true at all frequencies, whereas neutralization only gives compensation over a narrow frequency range. Neutralization however is much simpler and is the method invariably used. A basic treatment of this problem along with the more common neutralization circuits has been given elsewhere[4]. At its simplest, neutralization involves a capacitor or inductor suitably connected between output and input, the value of the capacitor or inductor being adjusted to cancel the internal feedback.

The accurate adjustment of the neutralizing component is a time-consuming operation requiring the connexion of instruments to the input and output of each transistor in turn. As a result it is common practice to use 'fixed neutralization' in which the value of capacitor or inductor used is that required to give exact neutralization on a 'typical' transistor. This approach is often satisfactory at lower frequencies or with lower stage gains. A major disadvantage however of any form of neutralization is that if gain control is applied to an amplifier stage the value of neutralizing component required varies with the transistor parameters which change with, for example, the emitter current and the collector-emitter voltage. This can lead to instability at certain gain settings.

The Cascode Circuit

An analysis of the cascode circuit is given by James[5], and a typical circuit diagram is shown in Fig. 1. The

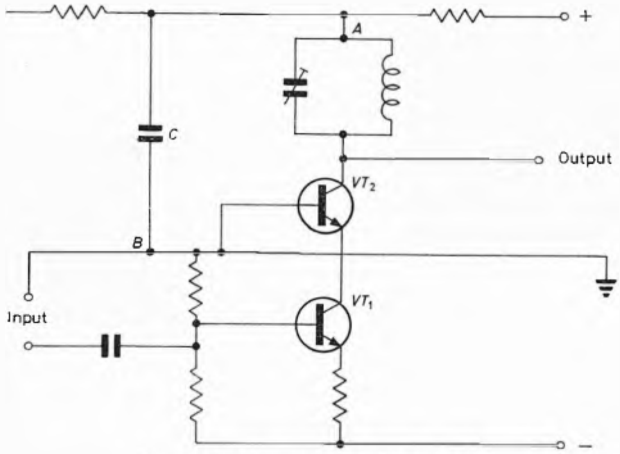


Fig. 1. Typical cascode amplifier circuit

circuit consists of a grounded emitter stage VT_1 whose collector load is a grounded base stage VT_2, the output being taken from the collector of VT_2. Because the (emitter) input impedance of VT_2 is very low, this collector-to-emitter connexion is a gross mismatch, and any change to the load of VT_2 has negligible effect on the input impedance of VT_1 (and vice-versa, any change to the source impedance has negligible effect on the output impedance of VT_2). This fact, and the fact that the internal feedback from output to input of the cascode is only 0·01 or 0·02 of that of a single stage is shown theoretically[5] by considering the parameters of the composite cascode stage.

In general a cascode can be used in place of a neutralized single stage amplifier without any of the latter's disadvantages. In addition it has two advantages: the cascode gain can be made greater than that of a single neutralized stage, and gain control can be applied with very little effect on the input and output impedances and on the stability. Several methods of varying the gain of a cascode amplifier have been described[6,7].

The Complementary Cascode

One of the principal difficulties encountered in building a multi-stage tuned amplifier is that of obtaining overall stability. This problem is independent of the stability of individual stages and is entirely a matter of eliminating unwanted feedback between the various stages. This feed-

* Ferranti Ltd.

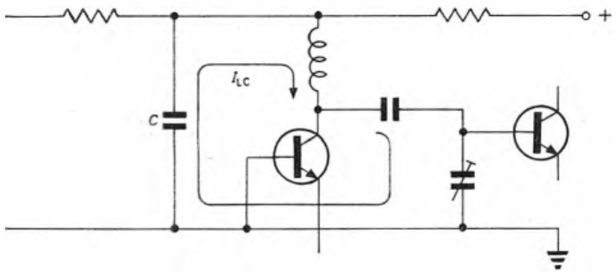


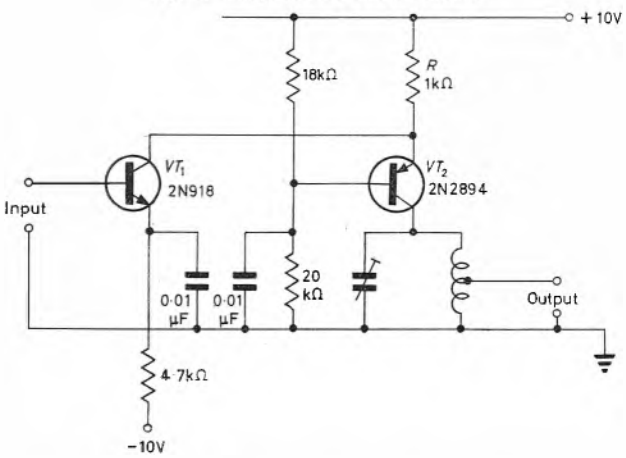
Fig. 2. Large I_{LC} current loop which results if tuning coil and capacitor are separated

back can be of two forms: radiated and conducted. Radiated feedback can be either electrostatic or electromagnetic while the conducted feedback is normally associated with the power supply rails.

Radiated feedback between stages can be eliminated by screening each stage from the next, and by screening the input and output of each stage from one another. Because this degree of screening is bulky, and in fact unnecessary, it is usual to compromise by screening the coils and by dividing the amplifier into two or three separate screened sections. (There is virtually no coupling between coils wound on screened pot cores). However it is normally overlooked that the decoupling capacitor, C in Fig. 1 is carrying r.f. and if the connexion between points A and B is of appreciable length it will form a very effective 'loop aerial'. If, as illustrated in Fig. 2, the tuning coil and capacitor are not connected directly in parallel, the circulating current I_{LC} (which equals Q times the input or resistive current to the tuned circuit) flows around an even larger loop. It is very important to make all r.f. connexions as short as possible, and to make the enclosed area of all current loops as small as possible.

Conducted feedback (due to r.f. signals flowing along the power supplies) is reduced by decoupling which should both prevent r.f. getting on to the supply rail from any of the stages, and prevent any r.f. which does get on to the supply rail from getting back into the signal path of an earlier stage of the amplifier. Decoupling shunts the signal currents along a low impedance path—a capacitor—by putting a high impedance in series with any other path. Unfortunately, at frequencies above 50Mc/s the self-inductance of a length of straight wire is appreciable[8]: at 100Mc/s a one inch length of wire can have an inductive impedance of about 15Ω. From this it follows that the actual amount of isolation may be very much

Fig. 3. Complementary cascode circuit



less than that calculated from the capacitive impedance alone. (Unless the capacitance is made to resonate with the lead length inductance[8]).

The cascode circuit shown in Fig. 3 tends to overcome these difficulties by virtue of its design. Comparing Figs. 1 and 3 it is seen that as far as the signal is concerned both circuits consist of a grounded emitter stage feeding into a grounded base stage. In Fig. 3 the direct current to both transistors is supplied through resistor R, whereas in Fig. 1 the same direct current flows through both transistors. Because the output tuned circuit is returned directly to earth the need for the decoupling capacitor C in Fig. 1 is eliminated, and the area of the signal current loop can be made much smaller, thus reducing radiated feedback. Resistor R is very much larger than the emitter input resistance of VT_2 and therefore decouples the r.f. from the positive supply rail.

There are three further advantages to be gained from the use of the complementary cascode. Firstly, because the collector of VT_2 is returned to earth through the coil it may be coupled directly to the input base of the following cascode. Elimination of the coupling capacitor reduces stray capacitance, electrostatic feedback and the recovery time after transient overload.

Secondly, consider varying the gain by varying the emitter current[6] in VT_1. When this is done (e.g. by varying the negative supply), with the complementary circuit, the total current drawn from the positive rail remains constant, and since there are no series decoupling resistors, the supply to all other stages remains at fixed potential. By the same argument, any change in the negative rail voltage has no effect on the current drawn from the positive rail.

Thirdly, if there are changes in the positive rail voltage, both the collector voltage and collector current in VT_2 will be increased (or decreased). The effect of these changes on the output impedance of VT_2 tend to cancel, thus reducing detuning of the tuned circuit.

Conclusions

An amplifier has been built at 80Mc/s using six stages of the complementary cascode to give an overall gain of 80dB and a bandwidth of 3Mc/s. This amplifier was built on double sided printed circuit using one side as a ground plane, coils were shielded pot cores, the positive supply rail was capacitively decoupled at either end of the amplifier, and no decoupling chokes or resistors were used in series with the supply. This amplifier is quite stable (with no evidence of unwanted feedback) when operated in the open—i.e. without being mounted in a screened chassis or box. This success is largely attributed to the complementary circuit which allows the r.f. currents to be confined to small area paths and eliminates the need for conventional power supply decoupling. In addition, the amplifier is very easy to align and there is no interaction between the tuning of different stages.

REFERENCES

1. Rexroad, W. D. Get Stability Through Mismatching. *Electronics*. *35*, 50 (19 Oct. 1962).
2. Jansson, L. E. High Frequency Amplification Using Junction Transistors. *Mullard Tech. Commun.* *3*, 174 (1957).
3. Stern, A. P. Aldridge, C. A., Chow, W. F. Internal Feedback and Neutralisation of Transistor Amplifiers. *Proc. Inst. Radio Engrs.* *43*, 838 (1955).
4. Wood, M. D. Gain Controlled Band-Pass Amplifiers (Part 1). *Electronic Engng.* *36*, 150 (1964).
5. James, J. R. Analysis of the Transistor Cascode Configuration. *Electronic Engng.* *32*, 44 (1960).
6. Perkins, J. F. Transistor Cascode Circuit Improves A.G.C. in Amplifiers. *Electronics*. *34*, 49 (2 June 1961).
7. Patent Specification 909,776. A Transistor Amplifier with A.G.C.
8. Lee, L. E. Rapid Selection of V.H.F. Bypass Capacitors. *Electronics*. *35*, 44 (28 Dec. 1962).

A D.C. Amplifier with Unity Voltage Gain

By G. G. Bloodworth*, M.A., A.M.I.E.R.E.

A direct-coupled buffer amplifier has been developed in which the voltage gain is very close to unity, with no offset between input and output, and there is first-order cancellation of the drift of transistor characteristics with temperature.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

A BUFFER stage is a very common requirement in instrument circuits. Such a stage must of course always have a high input resistance R_{in} and a low output resistance R_{out}. Very often direct coupling is needed and drift must be low. Finally, it is often convenient to have an incremental voltage gain very close to unity and also no step in voltage between input and output, so that the insertion of a buffer stage does not affect the analogue voltages but just provides isolation by virtue of the very large ratio R_{in}/R_{out}. A novel circuit has been developed which combines all these features. Although very few components are involved, the mode of action of the complete circuit may not be obvious, and so the development will be outlined in stages.

Grounded-Collector Amplifiers

The basic buffer amplifier is of course the grounded-collector stage shown in Fig. 1(a). For incremental changes, the voltage gain is less than but near to unity and a resistance transformation ratio is provided which is roughly equal to the current gain β. Thus the input resistance $R_{in} \simeq \beta R_L$, where R_L is the load resistance and the output resistance $R_{out} \simeq R_S/\beta$, where R_S is the source resistance. There is a step of about half a volt between input and output, given by the emitter bias voltage V_{eb}, and this voltage drifts with temperature, decreasing at a rate of about 2mV/°C with increasing temperature. Drift due to leakage current is usually less important ($R_S\Delta I_{CO} \ll \Delta V_{be}$) because ΔI_{CO} is minimized by using a silicon transistor, if R_S is not small.

The resistance transformation ratio can be increased by cascading grounded-collector stages, as shown in Fig. 1(b), to increase the overall current gain to the product of the individual current gains. The incremental voltage gain remains near to unity, but the voltage steps across the transistors are added together, and so are the corresponding drift components.

The emitter bias voltages can be made to subtract instead of add by using a complementary pair of transistors as shown in Fig. 2(a). The voltage steps and the drift components then cancel each other, at least to a first order. The voltage gain remains near to unity, but the current gain is reduced by the presence of the resistance R_1 which must be relatively low because it carries the standing currents from the emitter of VT_1 and the base of VT_2. Consequently, the resistance transformation ratio is reduced.

The current gain, and hence the transformation ratio, can be increased considerably without introducing another voltage step by connecting a third transistor as shown in Fig. 2(b). The combination of VT_2 and VT_3 is equivalent to a pnp transistor[1,2] connected as VT_2 in Fig. 2(a), with enhanced current gain roughly equal to the product of the individual current gains. To complete the circuit, the resistances R_2 and R_3 shown in Fig. 3 are needed to avoid passing bias currents through the source and load. The

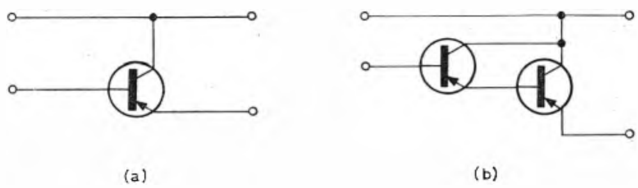


Fig. 1. Grounded collector and cascaded grounded collector stages

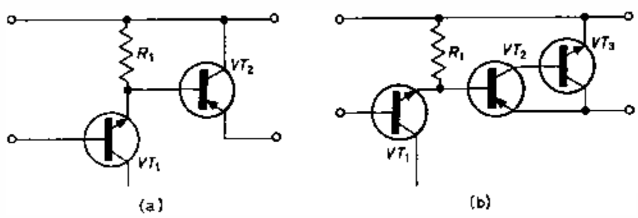


Fig. 2. Complementary pair and addition of third transistor

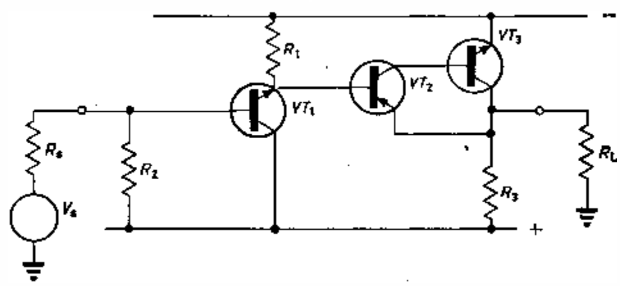


Fig. 3. Completed circuit with bias arrangements

input resistance of the amplifier is shunted by R_2, but the latter is large if VT_1 is operated at a very low current and reasonably high voltage supplies are used.

With zero input voltage the standing currents in VT_1 and VT_2 can be set so that the output voltage is also zero. A positive increment in the input voltage will cause the current in VT_1 to increase and that in VT_2 to decrease, and the changes in emitter bias voltage add together to give the amount by which the positive voltage at the output is less than that at the input. For a negative input voltage, all changes are reversed. The signal voltages

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across the emitter junctions always add together, and so the voltage gain must be less than unity by a significant amount, and further cascading of grounded-collector stages cannot alter this situation. It is however, possible by adding another transistor, shown as VT_4 in Fig. 4, to cause the currents in VT_1 and VT_2 to increase and decrease together so that there is no difference between input and output voltages, at least to a first order.

The Complete Amplifier

The transistor VT_4 provides the standing currents in VT_2 and VT_3 and it also acts as a grounded-emitter amplifier with its input resistance loading the collector of VT_1. As the phase reversals between base and collector in VT_1 and VT_4 cancel each other, a signal path is introduced which is parallel to the path through VT_1, VT_2 and VT_3. The collector of VT_4 provides a high resistance source of signal current which divides between the load and the output resistance of the upper channel, and as this output

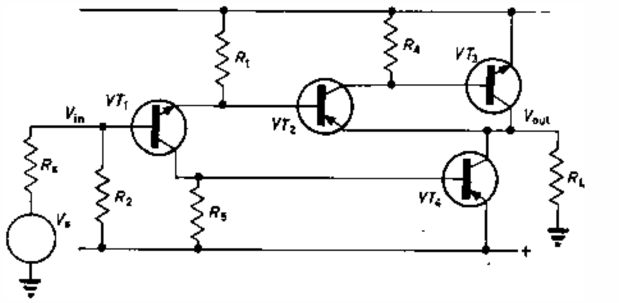


Fig. 4. Arrangement to avoid differences between input and output voltages

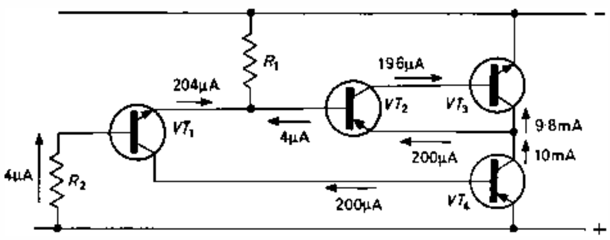


Fig. 5. Complete circuit with approximate currents

resistance is very low virtually no current passes into the load. The signal current flowing into the emitter of VT_2 opposes that caused directly by VT_1. In this way the incremental change in the emitter bias voltage of VT_2 can be reversed in direction and, by a proper choice of component values, made equal to that of VT_1 so that the overall voltage gain is increased to unity. This operation may be illustrated by an idealized example, as follows.

Suppose R_4, R_5 and R_L are infinitely large, and that for each transistor the collector leakage current is negligible and the current gain from base to collector is 50. With arbitrary voltage supplies R_2 can be chosen so that the base current of VT_1 is, say, 4μA and hence the collector current is 200μA. Thus the collector current of VT_4 becomes 10mA, and this divides between VT_2 and VT_3 as shown in Fig. 5. The sum of the base current of VT_2 and the emitter current of VT_1 determines the required value of R_1. It should be noted that the base current of VT_2 is proportional to the collector current of VT_4 and so is determined by the collector current of VT_1 but is not affected directly by the emitter current of VT_1. If the base current of VT_1 is increased by an input signal, all of the transistor currents will increase in the same proportion.

The approximate figures given in Fig. 5 show that VT_1 and VT_2 operate at roughly the same current and if complementary types are used the emitter bias voltages should cancel. The transistors VT_3 and VT_4 must clearly have equal current gains if the current in VT_2 is to be made equal to that in VT_1. In practice, of course, complementary pairs can only be roughly matched. However, if the effective current gain of VT_3 or VT_4 can be varied the voltages can be made to cancel even if neither pair is closely matched. The necessary control of current gain is provided by the resistances R_4 and R_5. As these resistances are linear but the input resistances of the transistors are non-linear, the effective current gains will vary with current level, even if the intrinsic gains do not, and so the current in VT_2 cannot follow that in VT_1 exactly. Also, the emitter characteristics of VT_1 and VT_2 will differ in shape and generate unequal voltage increments even with equal changes in current. Nevertheless, the accuracy of cancellation is very good in practice.

If the values of the resistors R_4 and R_5 are both decreased, without disturbing the essential balance of the circuit, then the standing currents in VT_3 and VT_4 are decreased relative to those in VT_1 and VT_2. This means that for a given input resistance to the amplifier less current is drawn from the battery, but the output resistance is increased because the current gain of VT_3 is effectively reduced. The standing current in VT_2 and VT_3 must of course be greater than the maximum current required by the load, but cannot exceed the input current multiplied by the current gains of VT_1 and VT_4. Within this limitation the circuit design is very flexible and tolerant to variations in characteristics between the four transistors. This will be illustrated by an example.

A Typical Design

Suppose now that the transistors are all different and that the various resistances in Fig. 4 are finite and are so chosen that V_{out} is zero if V_{in} is zero. If ΔV_{be1}, ΔI_{e1} and ΔI_{b2} are respectively the changes in the base-emitter voltage and emitter current of VT_1 and the base current of VT_2 produced by a change in V_{in}, then the input resistance of the amplifier is given by:

$$1/R_{in} = (1/R_2) + \frac{\Delta I_{e1}}{(\beta_1+1)\left[\Delta V_{be1}+R_1(\Delta I_{e1}+\Delta I_{b2})\right]}$$

where β_1 is the current gain of VT_1.

In the ideal case examined earlier, $\Delta I_{b2} \ll \Delta I_{e1}$. This inequality is maintained with practical values although ΔI_{b2} is affected by R_L and by the inequality of transistor gains. Also, the voltage change across R_1 is much greater than ΔV_{be1}, and therefore:

$$1/R_{in} \simeq (1/R_2) + \frac{1}{(\beta_1+1)R_1}$$

For given voltage supplies, the input resistance is therefore determined almost entirely by VT_1, which should have high gain and must be operated at a low enough current to give the required value of R_{in}. The base current of VT_2 does not affect R_{in} unless it is comparable with I_{e1} and therefore much greater than the base current of VT_1 when the transistors are operated at the same base-emitter voltage. If the characteristics are such that the npn transistor must be operated at a higher current than the pnp transistor, then the transistor types should be exchanged throughout the circuit, and the voltage supplies reversed.

Suppose the voltage supplies are ± 10V and, in an extreme case, a poorly matched pair of silicon transistors give, for $V_{ce} = 10V$ and $V_{be} = 600mV$

VT_1 (npn) $I_{c1} = 110\mu A$ $\quad I_{b1} = 5\mu A$

VT_2 (pnp) $I_{c2} = 1400\mu A$ $\quad I_{b2} = 33\mu A$

Then

$$R_1 = \frac{10-0{\cdot}6}{110+5+33}\,\mathrm{M\Omega} = 64\mathrm{k\Omega}$$

$$R_2 = 10/5 = 2\mathrm{M\Omega}$$

Ignoring ΔI_{b2}, which is now a gross approximation, the input resistance is very roughly 1MΩ. For VT_4, the base current cannot exceed 110μA, which limits the collector current (with $V_{ce} = 10V$) to 7·5mA in a typical case. If the load current will not exceed 5mA, the current in VT_4 may be set at, say, 5·5mA by introducing the resistance R_5 with a value of 22kΩ, found from the input characteristics of VT_4.

The collector current of VT_3 is:

$$I_{c3} = I_{c4} - I_{c2} = 5{\cdot}5 - 1{\cdot}43 = 4{\cdot}07\mathrm{mA}$$

and hence, from typical characteristics, the base current I_{b3} is 120μA. From the difference $I_{c2} - I_{b3}$ which is carried by R_4, the required resistance is found to be 500Ω.

The output voltage has now been set to zero for zero input voltage, because there is no current in R_L; this condition is of course independent of the value of R_L. A positive increment in V_{in} causes a reduction in the current in R_2 and an increase in all other currents, and current flows into R_L. The latter current would have flowed into VT_2 and VT_3 if the output were open-circuit, and so the effect of R_L is to make the current in VT_2 increase less than that in VT_1 and so V_{eb2} increases less than V_{be1}. Thus the positive output voltage is slightly reduced by R_L, which confirms that the output resistance of the amplifier has a low positive value. On the other hand, if V_{in} is negative the currents within the amplifier all decrease and V_{eb2} decreases less than V_{be1} due to the load current.

Performance of the Amplifier

As the transistors are driven over a wide dynamic range small-signal equivalent circuits are not particularly useful in the analysis of this amplifier, and calculations of input and output resistances and voltage gain must depend on the static characteristics of the devices and involve graphical or numerical analysis. A satisfactory approach is to calculate the input voltage and current corresponding to various values of V_{out} and R_L; this has the advantage that the difference in current between VT_4 and the combination of VT_2 and VT_3 is known. The procedure is as follows:

(1) Select initial values of V_{out} and R_L.

(2) From the characteristics of VT_2 and VT_3 (with $V_{ce} = 10 + V_{out}$) and the value of R_4 (500Ω), plot I_{b2} and V_{eb2} against $I_{e2} + I_{c3}$.

(3) From the characteristics of VT_4 and VT_1 (with $V_{ce} = 10 - V_{out}$) and the value of R_5 (22kΩ), plot I_{e1} against I_{c4}.

(4) Use the relation $I_{c4} = I_{e2} + I_{c3} + V_{out}/R_L$ to shift the axis and so plot I_{e1} against $I_{e2} + I_{c3}$.

(5) $R_1(I_{e1} + I_{b2}) = V_{out} + 10 - V_{eb2}$. Add the curves for I_{e1} and I_{b2}, multiply by R_1 (64kΩ) and add the curve for V_{eb2}, and from the ordinate ($V_{out} + 10$) determine the actual operating value of $I_{e2} + I_{c3}$ and hence the values of I_{e1} and V_{eb2}.

(6) From the characteristics of VT_1, find I_{b1} and V_{be1}.

(7) Calculate the input voltage $V_{in} = V_{out} + V_{be1} - V_{eb2}$ and the input current $I_{b1} + (V_{in} - 10)/R_2$.

(8) Repeat for different values of V_{out} and so determine the input and voltage transfer characteristics. Repeat for different values of R_L. Determine R_{out} from the variation in V_{out} with V_{in} constant.

(9) Repeat for different temperatures.

TABLE 1

	$R_L \to \infty$		$R_L = 1K\Omega$	
V_{out}	5V	−5V	5V	−4
V_{in}	5·014V	−5.016V	5.020V	−4·027
Voltage gain	0·997	0·997	0·996	0·993
ΔI_{B1}	3·2μA	3·2μA	3·2μA	2·8μA
R_{in}	890kΩ	890kΩ	890kΩ	830kΩ

Hence R_{out} is 1·2Ω at 5V and 3·5Ω at −4V, assuming R_s to be low.

Some of the results obtained with the above circuit values are given in Table 1; these were confirmed by experiment. The resistances are large-signal (chord) values at the voltages given. The relevant curves are slightly non-linear and give somewhat better values for smaller voltages.

With $R_L = 1k\Omega$, the transistors VT_1 and VT_4 cut-off when the input voltage is about −5V, and current flows only from R_L into VT_2, VT_3, R_1 and R_4. With a finite load this condition must always arise at some input voltage less than the negative supply, and for any such amplifier an appropriate set of values of minimum R_L and maximum negative V_{in} can be calculated. As VT_1 approaches cut-off, but VT_2 does not, the voltage gain is affected because the V_{be} values no longer cancel, and the input resistance depends on the load because I_{b2} is not much less than I_{e1}. This degradation is already apparent with $R_L = 1k\Omega$ and $V_{in} = -4V$ in this case. For positive and small negative input voltages R_{in} is essentially determined only by VT_1, R_1 and R_2, and when R_L is changed the currents in VT_2 and VT_3 change but those in VT_1 and VT_4 do not. The maximum positive input voltage is determined either by the allowable power dissipation in VT_3 or by the positive supply voltage.

The most significant effect of variation in temperature is to destroy the initial balance and produce an offset voltage between input and output, even if the transistors are so mounted that they are subject to the same variation. As expected, the drift found has been about 0·1 to 0·3mV/°C, which is typical of balanced amplifiers using unmatched transistors in long-tailed pairs.

Conclusions

The performance of the circuit arrangement which has been described could clearly be improved by the use of matched complementary transistors with high gain. In any case, it is easy to determine, by calculation or experiment, the resistance values for initial balance with satisfactory bias currents. The circuit then provides all the essential characteristics of a buffer amplifier with an attractive economy in the number of components.

Acknowledgment

The author wishes to thank D. N. Axford and R. Ball who made measurements and calculations on the amplifier during project work in the Department of Electronics, University of Southampton.

REFERENCES

1. LIN, H. C. Quasi-Complementary Transistor Amplifier. *Electronics*, *29*, 173 (September 1956).
2. SMITH, D. A., LEWIS, T. M. A. A Buffer Stage for Piezo-electric Strain Gauges. *Electronic Engng.* *34*, 99 (1962).

Feedback and Noise Performance

By R. E. Bogner*, M.E.

The fact that the noise figure and the optimal source impedance for an amplifier are independent of feedback is established by a very simple, general method. The derivation suggests a convenient way of determining the effect on noise performance of the circuits used in the feedback paths. Examples of its application are given.

(Voir page 137 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 144)

IT is an established and useful general result that the noise figure and the optimal source impedance for an amplifier are substantially unaffected by the application of feedback. The analysis presented here facilitates an understanding of the effect of each part of the circuit, and shows the conditions required for the effect of feedback on noise performance to be negligible. The results are implicit in the sophisticated Haus-Adler theory[1] but are deduced readily by an application of Peterson's theorem[2] (Appendix) in a manner which will be more digestible to most engineers.

The form of Peterson's theorem which will be used here states that the performance of a noisy, linear two port network may be described completely by the addition of two noise generators to a noiseless network which is equivalent otherwise. It is convenient to consider two generators, of voltage E_n and current I_n at the input of the amplifier as shown in Fig. 1.

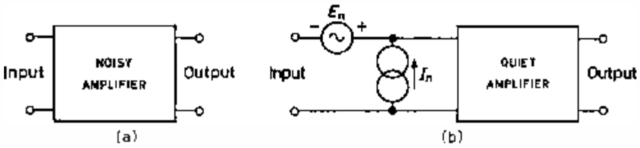


Fig. 1. Equivalent noise generators

Derivation

The noise performance of the amplifier may be determined by considering only the section containing the two generators, as the 'quiet amplifier' does not change the overall noise figure. The next step is to consider the application of feedback to the noisy amplifier as shown in Fig. 2(a). It is most convenient to consider all possible types of feedback and combining circuits at once to avoid numerous special cases. The system of Fig. 2(a) incorporates a noisy amplifier, defined by two noise generators, and a transmission matrix, $[A]$, such that:

$$\begin{bmatrix} E_o \\ I_o \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} E_1 + E_n \\ I_1 + I_n \end{bmatrix} \qquad \ldots\ldots\ldots\ldots (1)$$

where E_o, I_o are the output voltage and current,

E_1, I_1 are the *net* input voltage and current of the noisy amplifier

E_n, I_n are the equivalent noise voltage and current, referred to the input of the amplifier.

The net input, $\{E_1, I_1\}$ is derived linearly from the external input $\{E_i, I_i\}$ and the output, $\{E_o, I_o\}$ by the feedback-combining network, such that:

$$\begin{bmatrix} E_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} \\ \alpha_{21} & \alpha_{22} \end{bmatrix} \begin{bmatrix} E_i \\ I_i \end{bmatrix} + \begin{bmatrix} \beta_{11} & \beta_{12} \\ \beta_{21} & \beta_{22} \end{bmatrix} \begin{bmatrix} E_o \\ I_o \end{bmatrix} + \begin{bmatrix} E_a \\ I_a \end{bmatrix} \qquad \ldots\ldots\ldots\ldots (2)$$

where E_i, I_i are the external input voltage and current

$[\alpha]$ is the signal transmission matrix

$[\beta]$ is the feedback transmission matrix

E_a, I_a are the noise voltage and current generators representing the contributions of the feedback-combining network.

The 'measuring system' merely makes available to the feedback-combining network signals proportional to E_o and I_o.

Substituting for $\{E_1, I_1\}$ in equation (1) gives the equation of the system:

$$\begin{bmatrix} E_o \\ I_o \end{bmatrix} = [A] \left\{ [\alpha] \begin{bmatrix} E_i \\ I_i \end{bmatrix} + [\beta] \begin{bmatrix} E_o \\ I_o \end{bmatrix} + \begin{bmatrix} E_n \\ I_n \end{bmatrix} + \begin{bmatrix} E_a \\ I_a \end{bmatrix} \right\} \qquad \ldots\ldots\ldots\ldots (3)$$

It will be noticed that the output of the system, $\{E_o, I_o\}$ is assumed to be determined exclusively by the amplifier. In other words, the treatment ignores any feedforward through the β network which is reasonable when the gain of the amplifier is much greater than unity. This assumption also justifies the use of only two noise generators to represent the contribution of the feedback-combining network.

Next the effects of the noise generators $E_a + E_n$ and $I_a + I_n$ are represented by two generators, E_n' and I_n' at the input of the whole (two port) system. The noise performance of the whole system may be determined once these external generators are known. The equivalent system is described by the equation:

$$\begin{bmatrix} E_o \\ I_o \end{bmatrix} = [A] \left\{ [\alpha] \begin{bmatrix} E_i \\ I_i \end{bmatrix} + [\alpha] \begin{bmatrix} E_n' \\ I_n' \end{bmatrix} + [\beta] \begin{bmatrix} E_o \\ I_o \end{bmatrix} \right\} \qquad \ldots\ldots\ldots\ldots (4)$$

Equating (3) and (4) yields:

$$[\alpha] \begin{bmatrix} E_n' \\ I_n' \end{bmatrix} = \begin{bmatrix} E_n \\ I_n \end{bmatrix} + \begin{bmatrix} E_a \\ I_a \end{bmatrix} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (5)$$

Equation (5) shows that the equivalent generators E_n' and I_n' are independent of the feedback, or β coefficients. They depend only on transformations applied to the external signal, that is, on the α coefficients, and on the original amplifier.

The noise performance is thus exactly the same as if the wires connecting the feedback circuit to the output measuring system were 'earthed'. The noise generators E_n', I_n' may be compared directly with the signal. However, it is often more convenient to consider both signal and noise transformed by $[\alpha]$. This, of course, is equivalent to comparing the transformed signal with the original noise generators as shown in Fig. 3.

Important special cases of this theory are those where the 'noise' is in the form of d.c. drift—causing an error in output or in quiescent conditions. Bias stabilization is then directly related to the reduction of zero frequency gain by negative feedback, the bias drift referred to the input being unchanged.

* *University of Queensland, Australia.*

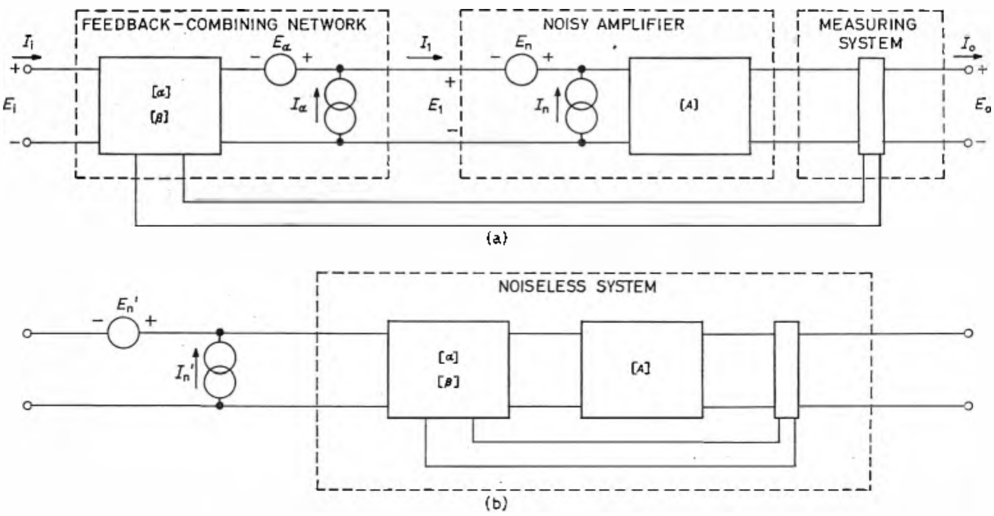


Fig. 2(a). General feedback system
(b). Equivalent system with noise generators replaced outside

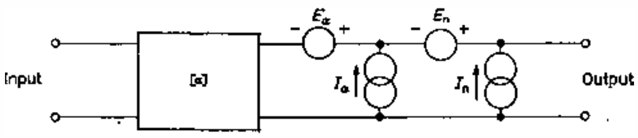


Fig. 3. Equivalent circuit for calculation of noise performance

The noise figure of the system, defined by:

$$F = \frac{\text{signal-to-noise ratio at input}}{\text{signal-to-noise ratio at output}}$$

is readily calculated from the two equivalent noise generators of a system[2] such as in Fig. 1(b). It is given by:

$$F = \frac{\overline{E_n^2} + 2p\,\overline{E_n I_n} + R_s^2\,\overline{I_n^2} + 1}{4R_s kTB} \qquad \ldots\ldots\ldots (6)$$

where $\overline{E_n^2}$ denotes the mean square of E_n

$\overline{E_n I_n}$ denotes the mean product $E_n I_n$

$\overline{I_n^2}$ denotes the mean square of I_n

where ρ is the correlation coefficient between E_n and I_n

R_s is the source resistance

kTB is the available thermal noise power, in the bandwidth B c/s appropriate to E_n and I_n.

To minimize F, with respect to variation of R_s, it is found that[2]

$$R_s = E_{n(rms)}/I_{n(rms)} \qquad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

Amplifiers in Cascade

The noise figure of an amplifier, defined above is not an adequate figure of merit in some cases. It is necessary to consider the gain as well. The classical expression for the noise figure of cascaded amplifiers is:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots\ldots\ \ldots\ldots (8)$$

where F_n is the noise figure of the n^{th} amplifier

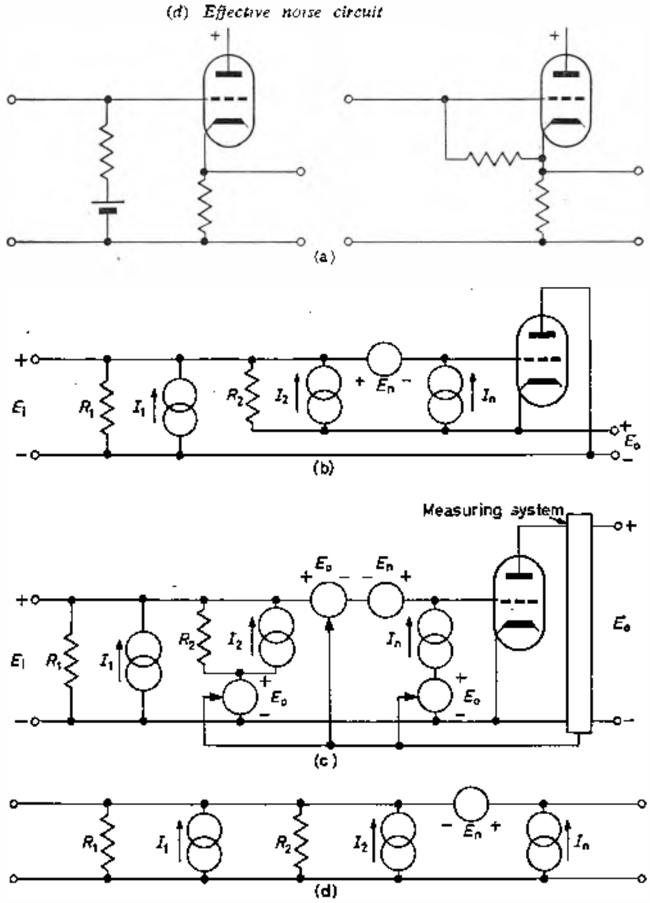
G_n is the power gain of the n^{th} amplifier.

If the G's are not large, then terms other than F_1 may make a significant contribution corresponding to the second or later amplifiers contributing significant noise. It has been shown that the noise figure of an amplifier is not affected by feedback over that amplifier. Thus, it is to be preferred to use little negative feedback over the first stage of a system, but to use overall feedback. In fact, it may be desirable sometimes even to consider positive feedback over the first stage(s).

Feed Forward

It is desirable to have some idea of the range of validity of the assumption that feedforward may be neglected.

Fig. 4. Cathode-follower
(a) Signal circuits represented by (b)
(b) General circuit of cathode-follower
(c) Effective feedback paths
(d) Effective noise circuit



Let a signal source deliver P_s watts of signal and P_n watts of noise to an amplifier. The internal amplifier noise is equivalent to N watts referred to the input. Let the amplifier and feedforward be such that the contributions to the output are:

gP_s watts of signal via amplifier

$g(P_n + N)$ watts of noise via amplifier

fP_s watts of signal via feedforward

fP_n watts of noise via feedforward.

Then the signal-to-noise ratio at the output is:

$$\frac{(g + f) P_s}{g(P_n + N) + fP_n}$$

If the feedforward factor, f, is neglected, the corresponding expression is:

$$\frac{P_s}{P_n + N}$$

The fractional error in neglecting f is then:

$$\frac{f}{g+f} \cdot \frac{N}{P_n + N} \leqslant f/(g + f) \leqslant (f/g)$$

Thus the fractional error in output signal-to-noise ratio is less than the fraction of the output determined by feedforward. Thus in just about any practical amplifier the neglect of feedforward is justified.

Examples

(1) Cathode-Follower

The method will be illustrated in clarifying the following points which have been discussed so frequently as to make the problem almost classical.

(a) The effect of noise sources within the valve.

(b) The effects of the grid return resistors in the two bias circuits of Fig. 4(a).

Consider the circuit of Fig. 4(b), which shows the essential signal circuit, including all significant noise sources. The resistors R_1 and R_2 represent the two types of biasing system—either resistor being infinite. The cathode-follower is considered to be derived from the grounded cathode amplifier by the subtraction of the output voltage, E_o, from the input voltage, E_i. The system is effectively as shown in Fig. 4(c). The principles explained above allow the significant features to be expressed as shown by Fig. 4(d):

(a) The noise generators corresponding to the valve (E_n, I_n) are effective directly at the input. From this it follows (equation (7)) that the optimal source impedance is the same as that for a grounded cathode stage with the same grid return resistor.

(b) The noise contribution of a grid return resistor of given noise current is independent of whether the resistor is returned to cathode or ground. The r.m.s. noise current associated with a resistor of R ohms is normally proportional to $1/\sqrt{R}$. Thus the preferred connexion for a given input impedance is to use the fixed bias circuit with the resistor grounded. (In this case, the resistor has a value $(1 + A)$ times that of the resistor required for connexion to cathode, where A is the voltage gain of the amplifier stage without feedback). The bias stability referred to the input is equivalent for both connexions if equal input resistances are to be obtained. This fact follows since grid current may be included in I_n, and cathode changes in E_n.

The implicit assumption that feedforward is negligible is amply justified under most conditions. Feedforward corresponds to coupling of signal via the grid-cathode impedance.

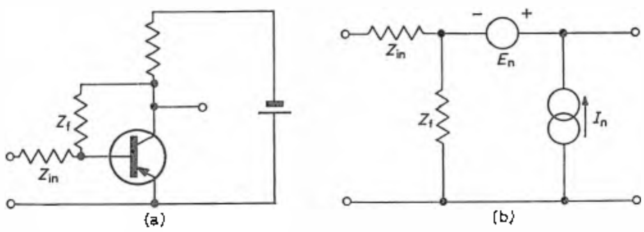


Fig. 5. Shunt feedback

The principle applies equally well to the emitter-follower, and in that case the saving in calculation over a conventional analysis is even greater.

(2) Shunt Feedback

Using the same ideas, the signal circuit of Fig. 5(a) is transformed to an effective noise circuit as in Fig. 5(b). The signal is transformed (usually attenuated) by the components Z_{in} and Z_f before comparison with the device generators. Of course Z_{in} and Z_f will often be associated with significant noise generators of their own, and these are readily included in the noise circuit.

Appendix

Representation of Noise in a Device by Equivalent Generators[3]

The following exposition concerns current and voltage generators referred to the input of a network. Similar arguments apply when the generator(s) are referred to other ports.

Referring to Fig. 1, it is required to make the two systems equivalent under all external conditions. The 'quiet amplifier' is defined as having the same transmission and impedance characteristics as the original amplifier.

Current generator I_n is chosen such that when the input is open-circuited, the open-circuit noise voltage appearing at the output is the same as in the original. I_n flowing in the input impedance of the quiet amplifier will cause a noise voltage to appear across its terminals, and to obtain the correct noise voltage across the external input terminals a generator of appropriate voltage, E_n, must be added. Thus, for the input circuit any external observations will be unable to distinguish between real and equivalent circuits (with the output open-circuited).

In particular, the change in input current and voltage when any external circuit is added is identical in both real and equivalent circuits. The corresponding change in output e.m.f. will thus be identical, since the transmission characteristics are the same. Hence, the output conditions must be identical whatever circuit is connected to the input.

It is now necessary to check that input conditions are satisfied for any output connexion. The output current and voltage change when any circuit is connected to the output are the same as in the original. Thus the change in any input circuit when an output circuit is connected must be the same as in the original, and it is known that the input conditions are identical for open-circuited output. Hence input conditions are always identical.

REFERENCES

1. Haus, H. A., Adler, R. B. Circuit Theory of Linear Noisy Networks. (Wiley, 1959).
2. Sanderson, A. E., Fulks, R. G. A Simplified Noise Theory and its Application to the Design of Low-Noise Amplifiers. *Trans. Inst. Radio Engrs. Audio*, p. 106 (July-August 1961).
3. Montgomery, H. C. Transistor Noise in Circuit Applications. *Proc. Inst. Radio Engrs. 40*, 1461 (1952).

Short News Items

A Symposium on the Applications of Microelectronics will be held at Southampton University from 21 to 23 September, 1965. This is being arranged jointly by the I.E.R.E. Southern Section and the I.E.E. Southern Electronics Section, in association with the Department of Electronics at the University. It is intended that papers, which will be pre-printed, will concentrate mainly on applications, although new circuit elements will also be of interest. Prospective authors are requested to send titles and summaries not exceeding about 300 words to the Symposium Secretary, Department of Electronics, University of Southampton, before 1 March, 1965. Registration forms will be available from the address from 1 April, 1965.

An international conference on thermionic electrical power generation is to be organized under the auspices of the O.E.C.D. European Nuclear Energy Agency by the Institution of Electrical Engineers. The conference, which is the first international meeting to be held in Europe on this topic will take place at the I.E.E., Savoy Place, London, during the week 20 to 24 September, 1965.

The object of the conference is to examine current research on thermionic processes for direct conversion of energy into electricity and to assist the dissemination of information on this new and promising technique, especially in its relation to nuclear and space applications.

Contributions are invited for inclusion in the programme. Those wishing to offer material are asked to send a provisional title and brief synopsis of their proposed contribution to the Conference Secretariat, (T.E.P.G.), The Institution of Electrical Engineers, Savoy Place, London, W.C.2.

The 6th International Conference on Medical Electronics and Biological Engineering (ICMBE&BE), sponsored by the Japan Society of Medical Electronics and Biological Engineering, under the auspices of the International Federation for Medical Electronics and Biological Engineering (IFME&BE), will be held in Tokyo, Japan, on 22 to 27 August, 1965, at Tokyo Prince Hotel. Topics to be dealt with at the Conference will be: the application of instrumentation, telemetering, television, computer, isotope, radiation, ultrasonics and optical techniques to clinical, basic and applied medicine and biology; and also mathematical analysis, simulation, bionics, biological control systems, artificial organs, ergonomics, human engineering, hospital automation and other problems related to medical electronics and biological engineering.

Further details are available from: Prof. K. Suhara, Secretary General, c/o Japan Society of Medical Electronics and Biological Engineering, Old Toden Building, 1-1 Shiba-Tamura-cho, Minato-ku, Tokyo, Japan.

The Fourth International Television Symposium, which is to be held at Montreux, Switzerland, on 24 to 28 May this year, will deal with the following subjects.

(1) Educational information on television—stressing the contributions in developing countries.
(2) Industrial and other application of television.
(3) Colour television.
(4) Television by satellites.

An international exhibition to be known as "Automatica 65" is to be held in Oslo, Norway, on 3 to 10 November.

This exhibition will be the fourth in a series with affiliated study conferences organized by the Norwegian Industries Development Association and will deal with instruments, apparatus and equipment for measurement control in engineering industries.

Further details are available from The Norwegian Industries Development Association at Forskningsveien 1-Oslo 3, Norway.

The British Radio Valve Manufacturers' Association (BVA) and Electronic Valve and Semi-conductor Manufacturers' Association (VASCA) state that the total sterling values of their members' sales of valves, tubes and semiconductor devices during the three months ending 30 September, 1964, were as follows:—Valves and tubes, £11·0M; semiconductor devices, £5·6M.

This brings the total value of sales for the first nine months of the year to £53·7M. The figure for the whole of 1963 was £61·6M.

EMI Ltd has produced a method of improving the signal-to-noise ratio of television pictures in order to keep this ratio as high as possible when dealing with camera tube signals.

It is necessary to employ a transmission channel whose bandwidth is sufficient to accommodate sharp transitions between black and white in monochrome transmission and in luminance in colour transmission. However, when the transition extends gradually over a comparatively large area of the picture, that is to say when the detail content of the picture is low, the signal bandwidth is much narrower than the channel bandwidth provided and an unnecessarily high noise to signal ratio results.

In a new method of overcoming this problem, the signals are divided according to their frequencies into a number of sub-channels by means of a number of band-pass filters whose characteristics cover a number of consecutive and coterminus frequency ranges which together cover the total bandwidth required. The output signal from each band-pass filter is applied to a threshold device, and the output signals from all the threshold devices are re-combined into a single transmission channel. The threshold devices may incorporate transistors or valves which are so biased that they are cut off when the level of the applied signal falls below the threshold value. Thus if the level of the signal in any frequency band is low, the output from the corresponding band-pass filter is cut off, so eliminating any noise in the cut-off frequency band.

In a colour television system, the band-pass filters and threshold devices may be included in the luminance channel to reduce luminance noise in the colour channel. The signal-to-noise ratio of audio signals may also be improved by the same method.

The fourth stage of the BBC's plans for extending and improving the coverage of the television service (BBC-1) and of the three sound services on v.h.f. by building further low-power relay stations in various parts of the country has received approval. The stations in this fourth group will, provided suitable sites can be found, be at the following places:

Relay stations for both television and v.h.f. sound

Scilly Isles	Dolgellau
Campbeltown	Llanidloes
Kingussie	Ballycastle
Lochgilphead	Kilkeel

Relay stations for television

Aldeburgh	Ballater
Bodmin	Girvan
Bude	Cardigan
Whitby	Llangollen
Ayr	Portrush

Relay station for v.h.f. sound

Weardale

A television station for Weardale is being built under Stage 3 and a v.h.f.

sound station already exists at Llangollen.

These stations will in general use 'translators' which receive the programmes from an existing station and relay them on another channel.

Stage 4 brings the total number of relay stations to be built by the BBC under its plans for extending and improving the coverage to 66 for television and 48 for v.h.f. sound. So far 27 of the television stations and 16 of the v.h.f. sound stations have been brought into service. When Stage 4 is completed, which is expected to be about the middle of 1966, the nominal coverage of BBC-1 will be 99·5 per cent of the population of the United Kingdom and that of the three programmes on v.h.f. sound will be 99·7 per cent.

Telefunken A.G. of West Germany has received an order from the Syrian Broadcasting Corporation for the supply of two 100kW m.f. broadcasting transmitters, which will be installed in Sabboura and Sarakeb and commissioned in 1965.

In addition the Company is installing near Djakarta a complete h.f. broadcasting station—fitted with a 100kW transmitter and antenna system—for the Indonesian Broadcasting Corporation. This transmitter will commence regular services at the beginning of this year.

The Company is also supplying a directional antenna system for a high-power transmitting station of the Kuwait Broadcasting System, comprising four h.f. transmitters with a total power output of 1 000kW.

Decca Radar Ltd has recently delivered its latest high performance weather surveillance radar type 43X to the University of Berlin. The equipment was installed by Telefunken A.G. of Ulm and is now fully operational.

The type 43X is an advanced 3cm weather radar employing a versatile aerial system to provide full facilities for range and height presentation of the vertical structure of precipitation. It is capable of performing advanced operational and research roles, and has replaced an earlier Decca meteorological radar at the Institut für Meteorologie und Geophysik der Freien Universitat Berlin which has been in service for several years and has provided meteorological information to the BEWAG weather bureau and to various other institutions.

With a range of 400km the Decca type 43X will give a picture of the weather situation over a very wide area, reaching to Scandinavia and the Baltic in the north, Austria in the south and Denmark and the Netherlands in the west.

An international consortium has recently been set up to compete for contracts worth £110M for a new NATO Air Defence System.

This system, known as NADGE (NATO Air Defence Ground Environment), will provide the most up-to-date forms of radar communications and data handling facilities, capable of giving the earliest possible warning and the fastest possible reaction to attack by supersonic aircraft. It will also speed up the feed-back of information and will provide a comprehensive integration of the NATO early warning chain, the Hawk/Nike missile complex and F-104 fighter squadrons in the NATO complex.

The members of this consortium, which combines unique experience in ultra-fast data handling techniques and a variety of new forms of radar designed to overcome jamming, are the Hughes Aircraft Company, Fullerton; Compagnie Francaise Thomson-Houston, Bagneux (Seine); The Marconi Company Limited, Chelmsford; Selenia S.p.A., Rome; Hollandse Signaal-Apparaten, Hengelo; and Telefunken A.G., Ulm. Conracts for this air defence system are expected to be awarded this year.

A scheme for the short-term hiring of up-to-date oscilloscopes has been arranged by Livingston Laboratories, 31 Camden Road, London, N.W.1.

The hire plan is limited at present to users within a 100-mile radius of London but will be extended as soon as practicable. Other electronic instruments will also become available.

Alberta Government Telephones, the Manitoba Telephone System, and Saskatchewan Government Telephones, have placed an order valued at £350 000 with AEI Telecommunications (Canada) Ltd for AEI centralized automatic toll ticketing (CATT) equipment and automatic number indentification (ANI) equipment.

The ANI equipment automatically identifies both the exchange and number of a caller using the Canadian direct distance dialling (STD in the U.K.) system, and also defines the class of call (business, private, etc). This information is passed automatically to toll centres where CATT translates this information, and provides extra data such as date, time and duration of call, and charging rate in a form suitable for assimilation by a computer or other business machinery.

The British UK.3 satellite, the third of a series in a joint United States/United Kingdom space programme is due to be launched from a site in the U.S.A. during the winter of 1966/67.

The satellite, which is to be built in the United Kingdom, will be designed by the Space Department of the Royal Aircraft Establishment, Farnborough, for a life of 12 months in orbit.

The satellite will carry five ionospheric research experiments. In general, each of the experiments is designed to produce information as a voltage varying between 0 and +5V. These voltages will then be sampled on a time multiplex basis via encoder circuits and the outputs applied to voltage controlled oscillators.

Construction of the data handling system which consists of three separate units comprising low speed encoder, high speed encoder and programmer, involves two systems of direct and recorded telemetry. The recorded information can be sent back at 48 times the recording speed on command from the ground. Data storage is by a specially designed tape recorder.

In the data handling system, a time-multiplexed equipment with a basic sampling rate of 55c/s, the voltage controlled oscillators are gated on and off to give bursts of about 10msec duration separated by 10msec blanks. The sampling sequence is arranged as a format of 6 frames of 16 channels for the replayed stored information. There are two operational modes for the low-speed format because the bandwidth requirements of the Jodrell Bank experiment are such that the whole of the recorder capacity is required. This will be done for one orbit in five.

The UK.3 satellite will be spin stabilized in a circular orbit which is expected to be at an altitude of 55km and inclined at 66°. It will be approximately 66in high and will weigh about 170lb.

G.E.C. (Electronics) Ltd, is the prime contractor for the electronic equipment and the British Aircraft Corporation is responsible for the manufacture of the satellite structure, assembly and testing of the complete satellite and for the ground test equipment.

The satellite will be assembled by the B.A.C., with assistance from G.E.C. in the integration of the electronic equipment and will be subjected to thorough environmental testing before it is sent to the U.S.A.

The M-O Valve Co. Ltd of Brook Green, London has received a £40 000 travelling wave tube contract from Lenkurt Electric Co. Inc. of San Carlos, California, a subsidiary of General Telephone & Electronics Corporation. The tubes will be installed in multi-channel radio communication networks in North America.

The contract covers the supply of M-O V. type TWC13 1 200-channel conduction cooled t.w.t.'s giving an output of 5W and a gain of 40dB over the frequency range 5 925 to 6 425Mc/s with an a.m./p.m. conversion of less than 2·5°/dB.

An automatic methane (firedamp) detector developed by the Mining and Materials Handling Division of The English Electric Co. Ltd has now received Ministry of Power approval for use in coalmines, and also in factories (in Class 2(c) pentane hazards). The detector continuously monitors the percentage of methane present in the surrounding atmosphere and provides a visual alarm when a preset danger level is exceeded.

It is a completely automatic, portable device suitable for operation by unskilled personnel.

Although in its present form the detector has been specifically designed for use in coalmines, it has applications in many other spheres such as the gas, oil, chemical and petrochemical industries.

The detector is self-contained and powered by a rechargeable lead-acid battery which is sufficient for ten hours continuous operation. In service, the methane-air mixture diffuses automatically to two heated sensing elements housed inside flame traps located at the top of the detector—making it possible to detect the presence of methane close to the roof of the mine workings. Each sensing element takes the form of a thermocouple sealed in a glass bead over which a heater coil is wound. One element—the detector element—is coated with a catalyst, while the other (compensating) element is untreated. The methane oxidizes on the coated element, causing a rise in temperature. The compensating element is unaffected, and the difference in thermocouple voltages is fed to a transistor switching amplifier, which controls the current to the omnidirectional alarm lamp. In a 60° window above the alarm lamp is a meter reading 0 to 3 per cent methane which is operated directly by the difference signal from the sensing elements.

Correction

In the article 'High Stability Low Frequency Band-Pass Filters' by P. G. Simpson, which appeared on pages 26 to 31 of the January, 1965 issue, several errors occurred in the equations given in the captions to the figures. The correct equations are as follows:

Fig. 2: $X = 2Q\,(1 - \gamma),\ \gamma = f/f_o < 1,\ Q \geqslant 100$

$$E = -20 \log \frac{Q^2\gamma^2}{2k/k_c}\left[\left\{-\frac{(k/k_c)^2}{Q^2} + 1/Q^2 - (1 - 1/\gamma^2)^2\right\}^2 + (4/Q^2)(1 - 1/\gamma^2)^2\right]^{1/2}$$

Fig. 3: $X = 2Q\,(\gamma - 1),\ \gamma = f/f_o > 1,\ Q \geqslant 100$

Fig. 6: $X = 2Q\,(\gamma - 1),\ \gamma = f/f_o > 1$ or $X = 2Q\,(1 - \gamma),\ \gamma = f/f_o < 1$

$$\theta = \tan^{-1}\left[\frac{2Q\,(1 - 1/\gamma^2)}{(k/k_c)^2 + 1 - Q^2\,(1 - 1/\gamma^2)^2}\right]$$

LETTERS TO THE EDITOR

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

Photomultiplier Fatigue

DEAR SIR,—I am surprised that there is no reference to the phenomenon of photomultiplier fatigue in the article by Gordy, Hasenpusch and Sieber[1].

Photomultiplier fatigue results in the anode current decreasing with time when the illumination and h.t. are constant. Typical decreases are 50 per cent in 15min. An almost complete recovery is made if the tube is stored in the dark for a few days. An example of this effect can be seen in Specification CV.5963 which is for a photomultiplier with caesium-antimony electrode surfaces, i.e. similar to the 1P21. This specification illustrates the measures required to counter fatigue effects when the tube is used as a noise source.

If photomultiplier fatigue occurs in the densitometer, the slope of the line relating V-V_o to optical density will be continually changing.

Would the authors care to comment on this?

Yours faithfully,

R. J. THRESHER,

The General Electric Co. Ltd, Stanmore, Middlesex.

REFERENCE

1. GORDY, E., HASENPUSCH, P., SIEBER, G. F. A Four Decade Transistor Linear Densitometer. *Electronic Engng.* 36, 808 (1964).

The Authors' Reply:

DEAR SIR,—With reference to the question of photomultiplier fatigue posed by Mr. J. F. Thresher, may I state simply that this phenomenon does not manifest itself at the 10^{-6}A photomultiplier anode current level which we use. Our own experience is limited in this respect to some five units. Personal communication with the chief engineer of a firm which manufactures photomultiplier devices reveals that in his experience, which covers a few thousand units operating in the field, fatigue is not experienced at these low photomultiplier anode current levels.

Yours faithfully,

EDWIN GORDY,

Roswell Park Memorial Institute, Buffalo, New York.

Simple Tuneable RC Null Networks

DEAR SIR,—I have read the letter published by T. C. Penn in the December issue (*Electronic Engng.* 36, 849, 1964).

The relation given for the Hall tuneable circuit (Fig. 3):

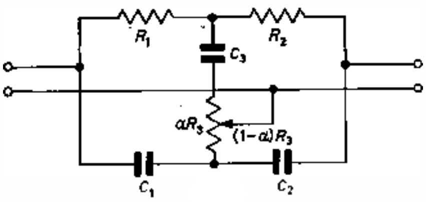
$$\omega_o = 1/[RC\sqrt{\{\alpha(1-\alpha)\}}]$$

is incorrect. It should read:

$$\omega_o = 1/[RC\sqrt{\{(1+2K)\,\alpha\,(1-\alpha)\}}]$$

The same false relation has been quoted in the article published by H. P. Hall in *The General Radio Experimenter*, 35, No. 7, July 1961, but the correct relation given above has been given in an erratum published in the October 1961 issue (*The General Radio Exp.*, 36, No. 10.

Fig. 1. Andreyev's circuit



When $K = 1$, the null condition is given[1] by the relation:

$$\omega_o = 1/[RC\sqrt{\{3\,\alpha\,(1-\alpha)\}}]$$

The letter published by T. C. Penn gives also the null condition for the Yu. A. Andreyev tuneable circuit (Fig. 2, of Penn's letter).

The general null conditions[2] for the circuit of Fig. 1 are:

$$\omega_o^2 = 1/[R_3^2C_3^2(1-\alpha^2)]$$

with the relations:

$$(C_1 + C_2) = C_3,\ C_1C_2/C_3^2 = R_3/(R_1 + R_2),\ R_1R_2 = R_3\,(R_1 + R_2).$$

Yours faithfully,

LEÓN GRILLET

(Maître de Conférences Honoraire, Faculté des Sciences de Rennes, France).

REFERENCES

1. GRILLETT, L., *Bull. Soc. sci. Bretagne*, 36, 307 (1961), and 37, 313 (1962).
2. GRILLET, L., *Bull. Soc. sci. Bretagne*, 36, 291 (1961), and 37, 91 (1962).

The correspondent replies:

DEAR SIR,—I commend León Grillet for bringing forth a more recent reference for H. P. Hall's tuneable *RC* null network which includes Hall's correction concerning the tuning relationship.

Concerning Andreyev's circuit, the tuning relationship previously published is correct and is a version of the more common symmetrical parallel-T. The previous circuit corresponds to León Grillet's version of Andreyev's circuit for $C_1 = C_2 = 2C_3$ and $R_1 = R_2 = 2R_3$. The more general null expression is useful however, to those who may wish to taper sections because of impedance relationships or because of a desire to change the null sharpness.

Yours faithfully,

T. C. PENN,

Physics Research Laboratory,
Texas Instruments Inc.,
Dallas, Texas, U.S.A.

A Novel Integrator

DEAR SIR,—Simultaneous development is no new thing in electronics, and an interesting example of it is the complementary transistor integrator described by Mr. Bertoya in your November issue. For a totally different purpose—that of respiration recording—we developed a very similar circuit and find it extremely convenient, as the input can be fed directly from a differential micromanometer, and the output connected (with suitable emitter-follower stages) to a meter or low-gain pen recorder.

One theoretical limitation is that, in any transistor we have so far looked at, the collector current is not entirely independent of collector voltage, but rises slightly over the range $V_{cb} = 1V$ to $V_{cb} = 10V$. This means in practice that, if the two transistors are exactly balanced, the capacitor voltage will always tend to drift slowly to a point midway between the voltages of the two collectors (or, if not quite balanced, some adjacent point).

This is rarely a disadvantage in practice, however—at least for integration of oscillatory signals—as the drift time-constant can readily be made long compared with the period of the signal being integrated: indeed, insofar as it helps to prevent the trace gradually drifting off the paper, it has proved a marked asset, particularly in clinical work.

Yours faithfully,

J. R. GREER,

Mercury Electronics (Scotland) Ltd,
Glasgow.

The Author Replies:

DEAR SIR,—Both Dr. Greer and your readers may be interested in yet two other ways in which the circuit has been applied at SIRA.

The first way concerned the maintenance of constant acceleration in a vibration test rig. The output from a sweep oscillator was fed through a variable $-\mu$ pentode via a power amplifier to the vibrator. The output from the accelerometer was taken via an amplifier to the integrator/comparator and the voltage across the integrating capacitor used as the control voltage for the variable $-\mu$ pentode.

By this means a change of 26dB in the magnitude of acceleration was compressed to within 0·5dB.

The second way was concerned with the measurement of optical transfer characteristic. A d.c. motor is used to alter the disposition of an optical element and so produce a signal which forms the input function to a lens. The output of the lens and hence its effect on the input function is measured automatically by a servo-system. In order that the system may operate at maximum speed, yet without exceeding the speed of response of the system, use is made of an existing tachometer. An additional output is taken from the tachometer to the integrator/comparator and the voltage across the integrating capacitor is taken via a series of emitter-followers to the d.c. motor. Thus, if the rate of change of output function tends to increase, say, the motor slows and the servo operates at approximately a constant speed.

No investigation has been made at SIRA of the effect of variation in collector voltage and unbalance in transistor pairs. However, like Dr. Greer, we have not experienced any trouble from these sources in the applications to date.

Yours faithfully,

H. C. BERTOYA,

British Scientific Instrument
Research Association,
Chislehurst, Kent.

An L.F. Revolution Counter and Frequency Meter

DEAR SIR,— Recently looking through the volumes of *Electronic Engineering*, I came across on page 395 in the June issue 1963, the article "A low frequency revolution counter and frequency meter" and in the September issue in the letters column a letter about a similar meter instrument. The circuit and details below show my approach to a related problem.

In the early part of 1962, I was consulted by a client in this country, who manufactures traffic indicator flashers for automobiles, and his problem was to regulate, align and test these flashers at a reasonable speed and produce units of closer quality tolerance. Up to that time the method employed was the time consuming test by counting flashes per minute or fraction of a minute. The method resulted in a large number of persons in the testing stage and above all a large part of the product was returned due to bad alignment and large variations of rate of flash.

The desired number of flashes per minute specified by various car manufacturers was from 65 to 75 flashes per minute. This is a slow rate of one per second. In order to test the flashers with the desired load in its natural conditions of operation, the unit was wired to a battery charged by a constant regulated supply. The pilot light normally mounted on the dash board was used to energize a photocell unit as in Fig. 1.

An OCP71 phototransistor in conjunction with an OC72 transistor VT_1 and VT_2 are connected in a monostable fashion, so that whenever the light level reaches or exceeds a certain point, the circuit changes state. When the light level falls below a certain value, the circuit returns to its original state. In order to minimize needle vibration at the read-out, and above all halve the time necessary for the needle to arrive at its final position, pulses from both the collectors of the OCP71 and OC72 monostable circuit are taken to charge a capacitor C by a transistor-diode pump circuit, via two amplifiers VT_3 and VT_4. Diode MR and VT_5 forms the pump circuit in which VT_5 improves the linearity of the charging characteristics. Thus this circuit pumps twice every time the pilot lamp flashes, that is when the monostable circuit flips and when it flops. VT_6 buffers the pump circuit. Potentiometer RV_1 adjusts the initial calibration of the meter, and RV_2 zeros the instrument.

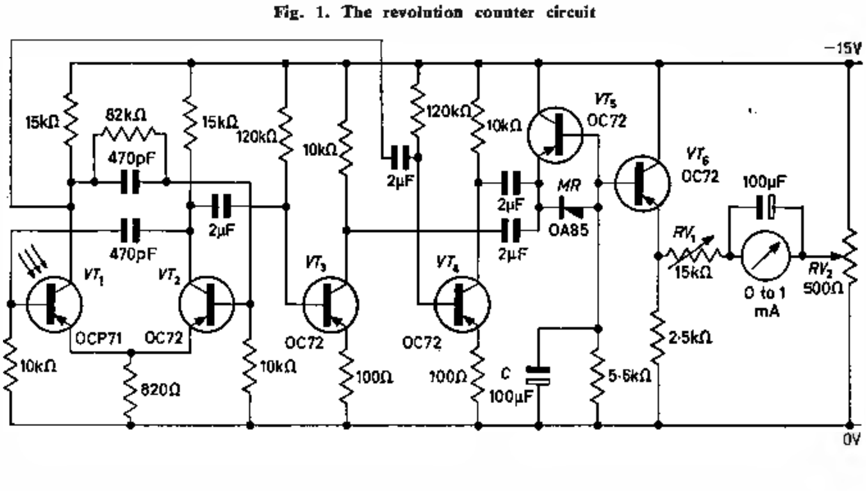
The result was that within 10 flashes the meter already indicated the final rate of flash in flashes per minute. The advantage of this circuit besides the above came to be that the flashes could be adjusted continuously on load until the needle falls within the tolerated limits. The time saved was immense. The circuit is slightly sensitive to the on-off time of the flash being unequal, but since alignment was done during working conditions continuously, the tester regulated easily to equalize the time on and off before proceeding to correct the rate.

Yours faithfully,

WET KUNG HSU,

Digimatic Ltda, Brazil.

Fig. 1. The revolution counter circuit



BOOK REVIEWS

Collision Phenomena in Ionized Gases

By Earl W. McDaniel. 775 pp. Med. 8vo. J. Wiley & Sons. 1964. Price 132s.

THE subject of electrical discharges in gases, or gaseous electronics as it is often called, has been a fruitful source of research for very many years. Moreover the fundamental processes are so many and varied that it is likely to continue to be so for some time. Recently there has been greatly increased interest in this field arising from, for example, magneto-hydrodynamics, controlled thermonuclear reactions, plasma guns, electron beam/plasma interactions, gas lasers, to name only a few of the new applications which are now possible.

In a comparatively few discharges, such as high-current arc plasmas, a condition of quasi-thermal equilibrium exists and the main discharge properties can be calculated (in principle at least) if the temperature is known. In general, however, this is not the case and a proper understanding of a discharge requires a knowledge of the various individual processes, both microscopic and macroscopic.

The book is largely concerned with collision phenomena in the broad sense, or what the author terms non-collective processes. It is designed apparently as a first-year graduate text and one is tempted to remark that the course concerned must be rather specialized and rather directed at research workers entering the field of gaseous electronics. In this sense the claim that the book is also useful for research specialists is justified. Considering the size of the work the author is to be congratulated on making it so up to date since references as recent as the end of 1963 are included.

The first chapter introduces the fundamental concepts of collision processes including the various cross-sections, collision probabilities and so on, and is followed by an outline of the principal relevant results of the kinetic theory of gases. Then follow two chapters, totalling over 100 pages, on the classical and quantum theories, and the measurements, of elastic scattering cross-sections. This topic takes up a large portion of the book and is quite comprehensively treated. A further large section (170 pages) is devoted to two chapters on inelastic collisions due both to electron impact and to heavy particle collision. The following chapter on photo-absorption in gases is quite short, perhaps a little too short, and contains a large number of curves of various absorption cross-sections. Some of these can be misleading in that experimental points are not shown (this also applies to other parts of the book) and the curves have been considerably smoothed, with the resonance lines and bands omitted. The next chapter on negative ions is followed by treatments of ion mobility, diffusion, electron energy distributions, drift, and recombination. The final chapter discusses surface effects including emission, adsorption and impact phenomena.

The book is clearly written, generously referenced and has excellent subject and author indexes. It is well produced, has a pleasing format and makes a useful addition to the existing literature.

W. A. GAMBLING

Fundamentals of Transmission Electron Microscopy

By R. D. Heidenreich. 413 pp. Med. 8vo. John Wiley & Sons. 1964. Price 109s.

THE aim of the book is defined by the author at the outset. It is "to develop a coherent physical approach to the understanding of contrast in electron microscope images, whether the objects are plant cell walls or crystals containing stacking faults."

The fundamental theory of all microscope imaging is due to Ernst Abbé. He regarded the object as a complex diffraction grating which acted upon the illuminating beam to create a diffraction pattern in the back focal plane of the objective lens; and this pattern, in turn, can be regarded as the source responsible for the final image produced by the microscope. The analysis is, in principle, always applicable to electron microscopy; and it is essential when crystalline objects are studied, and when the resolving power of the microscope approaches a few Angstrom units.

In the first two chapters the author summarizes the results that can be obtained neglecting the wave nature of the electron, and he exposes the limitations of this approach. In the bulk of the book he applies Abbé's theory to electron microscopes. Electron diffraction is considered in some detail—simple kinematic and two beam dynamical theory is explained. The contrast observed in actual microscope images is discussed in two chapters. Chapter IX is devoted to crystals, to examining how thickness variations, strains and lattice irregularities are made visible; Chapter X deals with image quality and the possibility of resolving individual atoms.

The book makes considerable demands of the reader. The treatment is concise, and previous knowledge of scattering and diffraction theory is desirable. Thus, Chapter VII, although entitled Introduction to Electron Diffraction, could hardly serve the purpose. Electron microscopists studying plastic and carbon replicas, and many biological specimens, can interpret their images adequately without referring to diffraction effects, and it is chiefly those concerned with crystal lattices and fundamentals of electron imaging who will find this work interesting and valuable.

The book is well produced—drawings and plates are excellent. There are some 350 references, mostly to recent work, and a set of appendices which contain useful physical data.

D. JONES

Radiation and Noise in Quantum Electronics

By William Louisell. 300 pp. Med. 8vo. McGraw Hill Book Co. 1964. Price £5 16s.

THE development of microwave and optical masers during the last decade has seen the introduction of a new term, quantum electronics. Broadly speaking, quantum electronics covers the interaction of electromagnetic radiation and matter. In some instances, this interaction can be treated adequately in purely classical terms, but in order to understand the basic principles of the maser and laser and to carry out research on these devices it is necessary to have a working knowledge of quantum mechanics.

The average physics graduate will have been introduced to the basic ideas underlying quantum mechanics but it is unlikely that he will have gone much beyond the hydrogen atom and a qualitative picture of atomic spectra. This introduction will probably have been in terms of the Schrödinger wave equation which is strictly limited in its application. Although considerably more abstract, the Dirac formulation of quantum mechanics provides the key to more advanced work and is the starting point for this book.

Following an introduction to the Dirac approach in the first chapter, the author proceeds to describe the harmonic oscillator and the electron in quantum terms. The manipulation of the mathematical functions which have been introduced is discussed thoroughly in Chapter 3, and in the next two chapters the author introduces the quantization of the radiation field and the interaction of radiation with matter. A chapter is devoted to the statistical aspects of large quantized systems and is related to spontaneous emission and noise and the book ends with a few applications of quantum mechanics to practical systems, such as the maser and parametric amplifiers.

This book can be recommended to any graduate physicist or electrical engineer who has the mathematical background and the desire to use quantum mechanics. He will not find it light-reading and will need to devote a considerable amount of time to it if he is to absorb the contents. Each chapter ends with references for further reading and a set of searching exercises. Anyone prepared to work his way through this book will no longer be scared to tackle the problems he will meet in the field of quantum electronics.

Like most books emanating from the U.S.A., it is relatively expensive but it is hoped that it will not deter too many from purchasing this excellent book.

C. S. BROWN

Physics of Semiconductors

By J. L. Moll. 293 pp. Med. 8vo. McGraw Hill. 1964. Price 92s.

J. L. MOLL has written this book on the physics of semiconductors for graduate students of electrical engineering.

The quasi-classical concepts of electrons and holes are justified in three chapters, the first entitled 'Quantum Mechanical Concepts', the second 'Electrons in Crystals', and the third 'Conductivity of Solids'. This is followed by a discussion of scattering mechanisms and statistical distributions. The next chapter is on carrier generation and recombination and this is followed by chapters on the application of the concepts and equations thus far derived to pn junctions and junction transistors.

The next section of the book is concerned with the behaviour of carriers in high electric fields and starts with a discussion of phonon spectra and scattering mechanisms in low and moderate fields. This is followed by an extensive chapter on secondary ionization and another on tunnelling. The final chapter is on surfaces.

The treatment throughout the book is quite highly mathematical but a number of figures are introduced to assist the reader to follow the argument. The mathematics are well presented although there are occasional blemishes, such as when a parameter is not defined till a few pages after it is first used, and when vague reference to a previous chapter, rather than a specific equation, is used to justify a step in the reasoning.

The author succeeds in the task he sets himself of establishing a good framework in the parts of the subject that he treats. However, the concentration of attention on these topics could well cause the reader to obtain an unbalanced impression of the subject as a whole since very little attempt is made to indicate what has been omitted. The author's decision to leave out a wide range of topics including such things as thermo-electricity, acoustic amplification and injection luminescence could well be justified on the grounds of expediency, but the omission should have been stated. Since the book was written for the electrical engineering student a similar criticism could be made that there are chapters on diodes (junction and Schottky) and junction transistors but no mention is made of any other semiconductor device.

As long as the reader is aware of these limitations on the scope of the book he should find it a useful introduction to semiconductors and the chapters on impact ionization and tunnelling are quite thorough reviews of these subjects.

J. R. A. BEALE

Instrumentation Techniques in Nuclear Pulse Analysis

337 pp. Demy 4to. National Academy of Sciences. 1964. Price $5

This book contains the papers presented at a conference held at the U.S. Naval Postgraduate School, Monterey, California, in 1963 and sponsored by the National Academy of Sciences—National Research Council.

Über das räumliche Hören (On Stereophonic Hearing) Biologische Periodik als selbsterregte Schwingung (Biological Periodicity as Self-excited Oscillation)

By Prf. Dr.-Ing. V. Aschoff and Dr. med. J. Aschoff. 101 pp. Royal 8vo. Westdeutscher Verlag. 1964. Vol. AFG-N.138. Paper back DM. 11.-

The first of the two papers deals with factors other than the sound source which affect the sound stimulus on the ear in enclosed spaces. Their influence on the transmission method and vice versa is considered. The second paper deals with the physical periodicity of certain body functions and activities of certain animals affected or controlled by their environment. Low frequency changes due to the change from day to night have hitherto been considered as typical example of a forced oscillation. New experiments have proved them to be self-excited.

Matrizen (Matrices)

By Prf. Dr.-Ing. R. Zurmühl. 452 pp. Demy 8vo. Springer Verlag. 1964. DM. 36.-

That the fourth revised edition appears only three years after the third one is a measure attached to this work which has two aims. It is intended to combine in one volume the theoretical principles and the laws of matrix algebra besides the calculation rules as far as they are essential to the understanding of published works and to the successful completion of work undertaken by the reader. Smaller and larger calculated examples, illustrations, schemes and ALGOL programmes ensure close linking of the theoretical and numerical presentation. The concluding chapter deals with matrices in electronics, statistics, elastomechanics and vibration techniques to give an introduction in kind and range of application and to facilitate the understanding of relevant literature.

Annual Review in Automatic Programming Vol. 4

By R. Goodman. 263 pp. Royal 8vo. Pergamon Press. 1964. Price 70s.

The fourth volume of this series of Annual Reviews by the Automatic Programming Information Centre at Brighton College of Technology, contains papers dealing with automatic programming and deals with such subjects as ALGOL and GIER-ALGOL compilers, JOVIAL and a specific IBM system.

ELECTRONIC EQUIPMENT

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

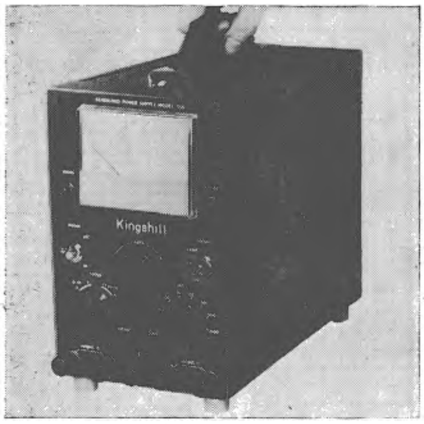
(Voir page 131 pour la traduction en français; Deutsche Übersetzung Seite 138)

STABILIZED POWER SUPPLY

Kingshill Electronic Products Ltd. Torrens Works, Torrens Street, London, E.C.1

(Illustrated below)

A compact laboratory model stabilized transistor power supply is now available from Kingshill Electronic Products Ltd, having an output of 0 to 50V at a maximum of 1A continuously variable in two switched ranges. The model 501 features comprehensive protection facilities for both source and load. The protection circuits, which are entirely electronic, can be set either to limit the load current or to cut off the output altogether at eight switched current values from 5mA to 1A. Selection of limit or cut-out function is by a slider



switch, and indication of overload is given by a lamp.

With regulation of 0·1 per cent, stability ratio of 1000, and ripple and noise less than 1mV this unit is rated at a maximum ambient temperature of 35°C. Each output leg has two parallel terminals which have 'castle' bases to prevent wire-straying. Output voltage and current may be accurately monitored on the large scale meter which reads 0 to 25V, 0 to 50V and 0 to 1A.

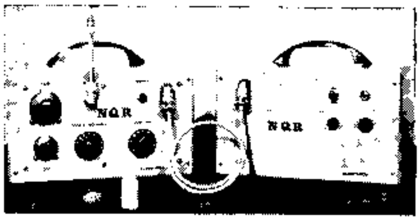
EE 78 751 *for further details*

NUCLEAR QUADRUPOLE RESONANCE SPECTROMETER

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

(Illustrated above right)

The Scientifica nuclear quadrupole resonance (n.q.r.) spectrometer has been developed specifically to meet the needs of science departments who require n.q.r. apparatus for both research and demonstration purposes. It is designed to operate over the frequency range of 10 to 60Mc/s and will detect many of the interesting nuclear quadrupole reson-



ances. Using the standard polycrystalline sample which is supplied along with the instrument, one may obtain large resonance signals from the two chlorine isotopes Cl^{35} and Cl^{37} at frequencies of 34·233 and 27·007Mc/s at a temperature of 25·2°C.

Measurements of the n.q.r. frequencies in samples containing more than one isotope yield a direct comparison of the nuclear quadrupole moments of the different isotopes, and, for some molecules, lead to the determination of their absolute values. For those nuclei whose quadrupole moment is already known, n.q.r. measurements make possible a prediction of the electron distribution inside the molecule, and this in turn provides information concerning the basic chemical structure of the sample. For example, they may enable one to decide whether a particular chemical bond is purely covalent, or purely ionic; or, for one of mixed character, what is the percentage contribution of each type.

Although in pure n.q.r. spectrometry, no magnetic field is necessary, it is possible to produce Zeeman splitting of the resonance line by the application of a weak field. Measurement of this splitting for different orientations of the sample in the magnetic field enables a calculation to be made of the degree to which the molecular electric field departs from axial symmetry. If an inhomogeneous magnetic field is applied to the sample, the resonance signal is smeared out by the line broadening which results. This effect may be demonstrated very easily by placing the small permanent magnet supplied with the kit within about four inches of the sample.

The general procedure for obtaining a nuclear quadrupole resonance signal is as follows. The sample is placed inside the coil of a detector oscillator, and the frequency of the oscillator is continuously varied. When it coincides with the value required for resonance, the sample absorbs energy from the oscillator, thus producing a change in the operating conditions of the oscillator detector circuit. It is arranged that the frequency is swept continually through the resonance condition, and this provides an absorption signal suitable for amplification by the low noise transistor amplifier incorporated in the circuit. The output from this may then be fed directly into a low gain oscilloscope for display.

In order to obtain the highest signal-to-noise ratio in the Scientifica n.q.r. spectrometer, a self-quenching oscillator detector circuit has been developed. It includes a control for altering the quench frequency, and this enables one to distinguish between the central resonance frequency and its side bands. Another control, marked 'MOD. AMP.', adjusts the width of band over which the oscillator frequency is being swept.

In addition to the rigid sample probe useful for n.q.r. measurements at room temperature, a flexible probe, twelve inches long, is supplied. This enables the sample to be placed in a cold temperature enclosure, or in a magnetic field for measurements of the Zeeman splitting. A length of silvered wire is also included for making up additional sample coils to meet individual requirements, either to accommodate samples of different sizes, or for coarse adjustment of the operating frequency.

EE 78 752 *for further details*

DIGITAL POTENTIOMETER

Scientific Furnishings Ltd, Poynton, Cheshire

(Illustrated below)

The new digital potentiometer developed by Scientific Furnishings Ltd combines rapid convenient single-knob null balancing with the advantages of digital indication of the measured voltage. The decimal point shifts automatically as the range is changed from a minimum of 0 to 9·99mV to 0 to 499·9mV maximum. A plug-in adaptor extends the range up to 500V with current ranges from 0 to 100μA up to 5mA. The measuring sensitivity is better than 0·2°C with an iron-constantan thermocouple or 10μV, with an absolute accuracy of 0·15 per cent.

An internal Weston standard cell is provided for rapid standardization and a 14Ah mercury battery provides a pre-

cision voltage source up to 1 000V for general calibration and test purposes. A leather camera-type case is available for field measurements for which the compact unit (5¼in by 5¼in by 7¾in) is particularly suited. The shock resistant galvanometer is protected from overload by a semiconductor circuit.

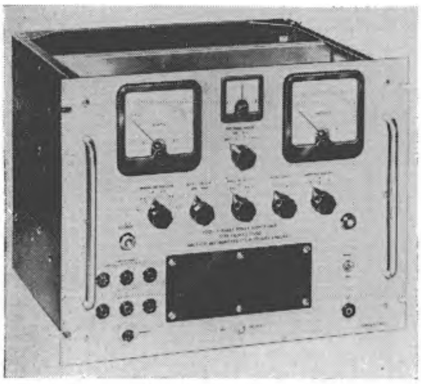
EE 78 753 *for further details*

400c/s POWER SUPPLY

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devon

(Illustrated below)

A new, three-phase, transistorized 400c/s a.c. power supply has been developed by Hatfield Instruments Ltd. It is designed primarily for the use of precision gyroscope manufacturers. The 400c/s oscillator employed in this unit is of the multivibrator type with a stability of ± 1 per cent and may be locked to an external source if required. A phase meter is provided which is substantially independent of voltage. This greatly facilitates the precise setting up of the 120° displacement between the three phases. Output voltage stability of 0·2 per cent is achieved.



EE 78 754 *for further details*

INDUSTRIAL CONNECTORS

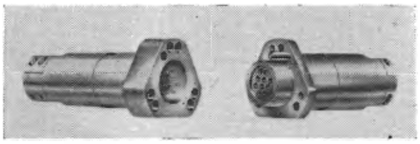
Plessey-UK Ltd, Ilford, Essex

(Illustrated above right)

The general trend towards automation, instrumentation and remote control in industry, particularly in the heavy industries such as mining and petroleum, increases the need for reliable, robust electrical connectors, which can accommodate a number of low power and signal circuits.

The Plessey-UK mark II and IIF ranges of connectors, couplers and cable and conduit fittings have been developed in close liaison with the National Coal Board to meet these requirements with maximum reliability and environmental resistance, and minimum size and cost. They comply with the N.C.B. specification for multiway connectors and are being adopted as their standard fitting. They are currently used on R.O.L.F. and Collins Miner installations and on conveyor and telephone systems in N.C.B. mines.

The mark II design is based on the



Plessey UK-AN connectors, hundreds of thousands of which are giving satisfactory service under arduous conditions. The same high quality silver or gold plated brass contacts for crimped or soldered conductor joints, and the same moulded silicone rubber insulators, are used in specially designed bolted flanged housings, made from high tensile brass forgings or bar. This design of flange permits plugs and sockets to be panel mounted or used as free items with or without cable glands or conduit fittings attached.

The four basic sizes of connector house various contacts rated up to 12, 24, 50 and 100A ranging from 2 to 48 ways. Polarizing keys and keyways prevent mismating, and cross-mating of similar connectors is prevented by using orientated connectors. These have six threaded holes in each flange into which the user may screw three orientation pins in any of 20 combinations so that plugs will only mate with complementary sockets.

Glands which screw on to plugs or sockets are specially designed to accommodate a large range of diameters of armoured or non-armoured cables in each basic size. They waterproof, seal and separately clamp on the outer cable cover and clamp the armouring where applicable. Conduit fittings clamp flexible conduits and conductors to the connectors.

The mark IIF (illustrated) range of connectors and glands for armoured and non-armoured cables is the flameproof version of the mark II. They comply with the requirements of BS 229 for flameproof electrical apparatus and are covered by Buxton Certificate Nos. FLP 5344 and 5345 for use with groups I, II and III gases.

EE 78 755 *for further details*

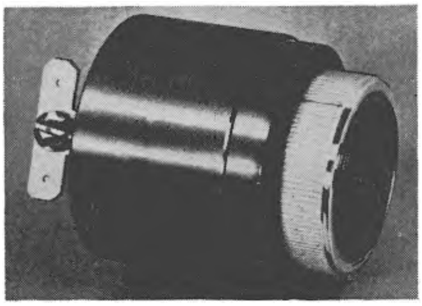
AUDIBLE INDICATORS

Highland Electronics Ltd, 26-28 Underwood Street, London, N.1.

(Illustrated below)

The Sonalert d.c. solid state audible indicators are now available in a new mains 200 to 250V a.c. version.

These devices give an obtrusive, easily identifiable, high frequency audible warning signal. They are small,



lightweight, self-contained with one hole fixing. They contain no moving contacts, generate no electrical interference, have high reliability and are designed for difficult environmental conditions. The size is 1·11/16in diameter by 2¼in long.

These Sonalerts have a very wide range of applications for announcing, signalling, monitoring, alerting and warning.

EE 78 756 *for further details*

FEEDBACK RESOLVERS

Moore, Reed & Co. Ltd, Woodman Works, Durnsford Road, London, S.W.19

(Illustrated below)

A new 400c/s, size 11 compensated resolver, series 11RSF, for precision computing by Moore, Reed and Co Ltd, is now in quantity production.

Deviation from sine wave of under 0·05 per cent with an inter-axis error of less than 4min is claimed for this resolver which, in production, has achieved a remarkably high proportion of units with deviations of under 0·03 per cent and with fundamental noise levels of 1 to 3mV. The transformation ratio of stator and compensation windings is 0·97.

Despite their high accuracy and general

quality, the units are low priced and are available with flying leads or with a 12-way terminal block.

EE 78 757 *for further details*

VARIABLE PHASE OSCILLATOR

Feedback Ltd, Crowborough, Sussex

(Illustrated on page 126)

The variable phase oscillator VPO230 is an example of operational amplifier techniques applied to a wide range sine wave generator.

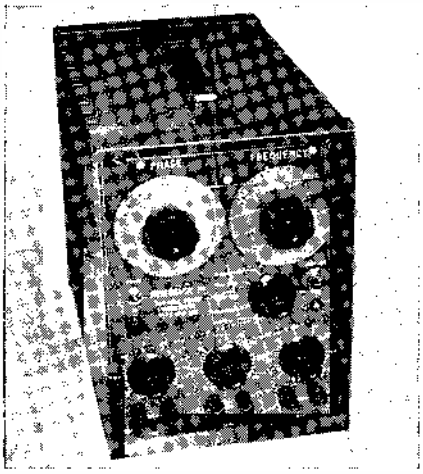
Particular features of this instrument are its three separately adjustable outputs—reference, quadrature and variable phase, Once set, these are at constant amplitude over the frequency range of lc/s to 100kc/s. Output impedances are 600Ω.

The Ref: Quad output is maintained within 0·25° of 90° and the variable phase may be set at any desired angle within 1°.

When operating as an oscillator, the outputs have a harmonic content less than 0·5 per cent.

The technique for phase and amplitude measurement involves the use of an oscilloscope with the variable phase signal applied to the horizontal deflexion channel.

The vertical channel is used to measure amplitude while the Lissajous display is brought to a null and phase



angle read directly from the phase dial on the oscillator.

The versatility is enormously increased by the external locking facility. This means that the instrument can be a 'slave' to any other frequency source within its range. Phase measurements are thus possible on systems operating at high power levels.

When operating under 'slave' conditions, the instrument behaves as a frequency selective amplifier and can be tuned to select harmonic components from a complex master signal.

EE 78 758 *for further details*

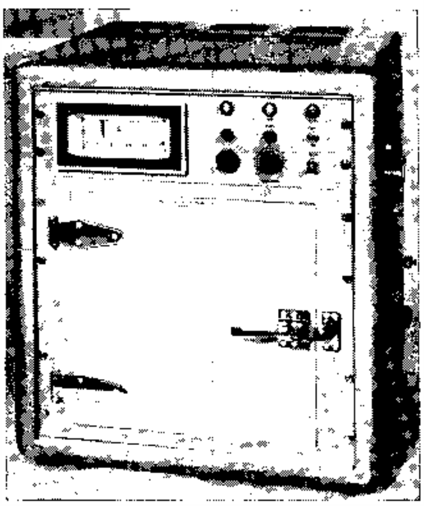
TEMPERATURE ENVIRONMENTAL TEST CHAMBERS

Montyford Instruments Ltd, 79 Lots Road, London, S.W.10

(Illustrated below)

The test chamber has a range from −70°C to 200°C and offers a rapid temperature change which can be controlled from the front panel. As well as variable rates of temperature change a hand-auto switch permits cycling between two preset temperatures or high stability at one temperature.

Programmes can be stopped or reversed at any desired time. Cooling employs CO_2 whose injection to the heat exchanger is governed by an electrical timer. Heating is incorporated in the



forced air flow circulation, eliminating hot spots and minimizing temperature gradients.

The chamber and control panel are modular built and are housed in standard width racking which can be extended to mount programme controllers, amplifiers and other rack mounted equipment which may be associated with equipment and components under the test.

Facilities for electrical connexions into the test chamber are provided making it suitable for the dynamic testing of components and equipment.

EE 78 759 *for further details*

TORQUE SENSING PLATFORM

Mechatronics (London) Ltd. 317 Kennington Road, London, S.E.11

A new type of torque transducer has been introduced by Mechatronics (London) Ltd.

Designed particularly for testing of f.h.p. motors, hydraulic pumps and motors, generators and alternators, the transducer is basically a torque measuring structure which eliminates the need for the conventional bearing of a reaction dynamometer.

It provides both the support for the power device and an electrical signal proportional to the reaction torque developed by that device. It consists of a plate supported from a fixed mounting base by means of torque-sensitive springs. Electrical strain gauges are bonded to these springs, forming a strain gauge bridge.

The springs are mounted at 45° angles. The intersection of a line drawn through the flexure springs describes the centre of rotation of the platform. Any device mounted to the platform with its input or output shaft precisely on this centre of rotation will produce a reaction torque which will be sensed by the platform.

Since there are no bearings or slip rings the torque platform has no friction loss and no speed limitation. This feature allows the torque platform to be used at very low torques without friction errors.

The bridge resistance is 100Ω and the bridge voltage is 12V; the output being 2mV/V nominal. Ranges available are from 0 to 3 lb.in to 0 to 300 lb.in.

EE 78 760 *for further details*

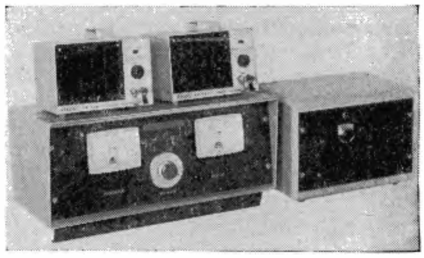
DIGITAL SERVO COMPARATOR

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

(Illustrated above right)

The function of this comparator is to compare two frequencies and produce an output proportional to their difference. This may take the form of pulses the frequency of which is the difference between the originals, or it may be an analogue voltage which is proportional to the frequency difference.

Basically the device first squares up the input signals in amplitude and time and then subtracts one complete pulse



of the higher frequency once per pulse of the lower frequency. This action is independent of the phase relationship between the two signals.

The output waveform is a series of pulses of width $\frac{1}{2}f_1$ and frequency $f_1 - f_2$. This is used as a digital output. The pulses are then averaged and smoothed. This is used as an analogue output.

The advantage of the arrangements is that there are no timing circuits involved, hence the accuracy is only dependent on the rise times of the output devices. The equipment may be made even more versatile by the addition of a few elements to make the device differential. In the first arrangement, if $f_2 > f_1$, the output is zero. In the latter arrangements the output is $f_1 \sim f_2$ together with an indication of which signal has the higher frequency. Digital display can be arranged merely by gating the outputs into a register in fixed time intervals. Alternatively a voltmeter can be used directly with the analogue output.

For applications such as automatic speed control, power amplification up to 2·5kW can be provided.

The fundamental frequency of the equipment is d.c. to 10Mc/s and the inputs may be between 3V and 100V; the input impedance is 10kΩ. The output is 6V at 1mA (digital) or 0 to 6V at 1mA (analogue).

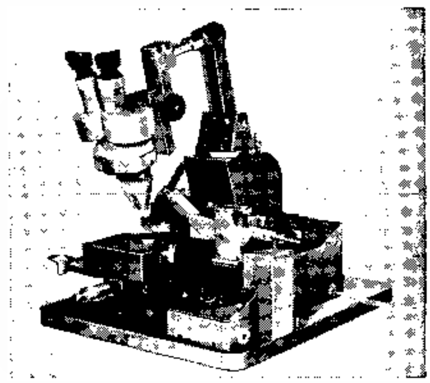
EE 78 761 *for further details*

SEMICONDUCTOR TEST PROBE

Vacwell Engineering Co. Ltd, Sherman Road, Bromley North, Kent

(Illustrated below)

A further new addition to their range of equipment for the semiconductor industry is announced by Vacwell Engineering Co. Ltd. Called the 'laboratory wafer prober' type PR.61, it is intended primarily for the spot checking of transistor dice while still in the wafer form,



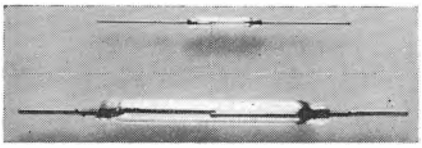
but has other uses, particularly in micro-circuit work.

The instrument is bench-mounted and is normally supplied with two probe heads, although provision is made for fitting as many as four. The probes are electrically insulated from the main body and are raised and lowered on to the wafer by operation of a small hand-lever. A separate, motorized inking probe, with push-button control, enables any faulty wafers to be marked *in situ*.

An insulated vacuum chuck, mounted on a micrometer-adjustable *X-Y* platform, holds the wafer undergoing examination and can itself be rotated to facilitate wafer alignment.

The specification of the PR.61, which can be readily coupled to a Vacwell semi-automatic three-parameter transistor tester or other test equipment, includes a 50-power magnifying binocular microscope and provision for illumination of the work platform. The microscope attachment is arranged to swing clear of the platform while initial placing of the wafer is being carried out.

EE 78 762 *for further details*



SUB-MINIATURE REED SWITCHES

Flight Refuelling Ltd, Wimborne, Dorset

(Illustrated above)

Flight Refuelling Ltd now has available the latest form 'A' reed switch from Hamlin. This new sub-miniature reed switch, the MTRG-2, is designed as a matching companion to the recently introduced MRG-DT form 'C'. As the MRG-DT and the MTRG-2 do not need magnetic biasing and as the contact gaps of both switches are in alignment, it is now possible to obtain double-pole changeover, single-pole changeover and single-pole single-throw combinations in the minimum size package.

Overall length is 2¼in with a glass diameter of 0·090in and the leads can be cut to 0·75in for packaging. These switches have inherited all the environmental characteristics of the standard size reed switches with life expectancy at full load of 3W d.c. resistive: 28V maximum, 0·250A maximum, of 10×10^6 operations, with contact resistance at 0·150Ω initial maximum and 0·100Ω nominal.

For a comparison of size the illustrations shows the MTRG-2 sub-miniature switch alongside the standard size from 'A' DRG-2.

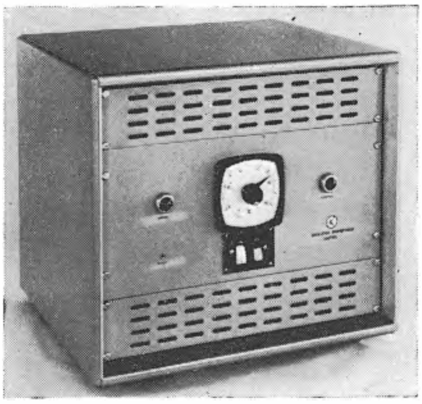
EE 78 763 *for further details*

TEMPERATURE CONTROLLERS

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated above right)

Kent Precision Electronics Ltd has produced a range of solid state temperature control instruments of advanced



design for control of electric furnace temperatures up to 2000°C with an accuracy of 0·5 per cent. The systems consists of conventional temperature transducer elements which feed a signal proportional to temperature into an accurate transistor d.c. amplifier. This input signal is compared with a stable and accurate reference signal, the magnitude of which is set as required on an indicating scale on the front of the instrument, and the error voltage is used to control the thyristor power output system to regulate the power delivered to the heating elements.

For low powers, proportional control of the output power is achieved by variation of the phase angle at which the thyristor circuit is fired during each half-cycle of the a.c. mains voltage. For high powers, to overcome interference, surges and poor power factors, the thyristor circuit is switched on for integral half-cycles, proportional control being achieved by variation of the ratio of the number of on and off cycles. In all cases proportional control is operational over approximately 3 per cent of the scale.

For control of currents over 100A when thyristor control becomes uneconomical, standard low power thyristor output circuits may be used to control saturable reactor elements for proportional control, or alternatively, electromagnetic contactors for on/off operation. The photograph shows a standard 10kW proportional temperature control unit.

EE 78 764 *for further details*

MINIATURE PRESET POTENTIOMETER

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

The type 62 is a new screwdriver-adjusted preset potentiometer made by Morganite Resistors Ltd. It is intended specifically for use in miniature equipments where small size is all-important. The component measures approximately 12 × 9mm.

Tags are spaced to suit standard 2·5mm (0·1in) module piercing. The contact can be adjusted from either side of the component and the resistance track is protected against accidental damage during adjustment.

Of the two versions offered, both plug directly into the printed circuit board. The type 62H mounts in an upright position.

The power rating is 0·2W at 40°C ambient and the resistance range is 100Ω to 1MΩ.

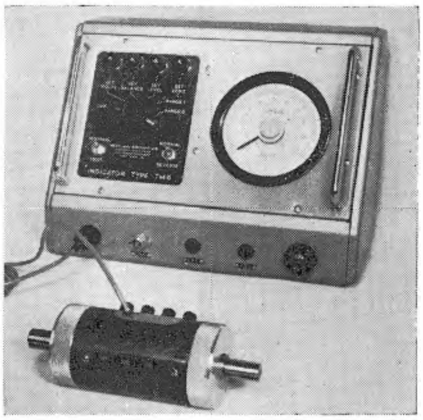
EE 78 765 *for further details*

TORQUE TRANSDUCERS

Saunders-Roe Division, Westland Aircraft Ltd, Yeovil, Somerset

(Illustrated below)

The new type 2 mark 4 torque transducers feature both rotor and stator assemblies in one-piece moulded fibreglass construction. Torque is transmitted through a high tensile steel shaft on which are mounted Saunders-Roe foil strain gauges. Electrical connexion between the rotor and stator is effected by silver sliprings and silver graphite brushes, and thence by a four-core integral cable to the instrumentation unit. The total errors of the transducer due to linearity, hysteresis and repeatability are less than ±0·2 per cent.



Instrumentation for this transducer is provided by the type TM6 indicator which gives two full scale torque ranges from any one transducer. Apart from the direct Cirscale read-out of the true mean torque, sockets are provided for the connexion of ancillary recorders, oscilloscopes and other optional additional equipment. The accuracy of true mean torque read-out is unaffected by large cyclic variations of torque.

The system operates from zero speed to the maximum rated according to the size of the transducer, and has a very high frequency response for transient and cyclic investigations. The indicator may be powered either by a.c. mains or battery supplies, and the transducers are available in seventeen ratings varying from 0 to 5 lb.ft up to 0 to 50 000 lb.ft, and speeds from 0 to 8 000rev/min to 0 to 800rev/min.

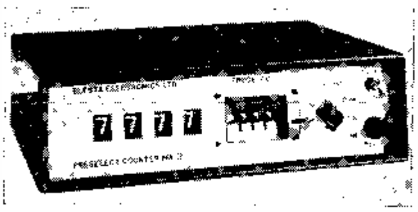
EE 78 766 *for urther details*

BATCHING COUNTER

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(Illustrated on page 128)

This compact, portable instrument provides for counting at up to 100kc/s. It uses printed circuit board construction



on a plug-in system and uses silicon devices throughout.

The provision of an edge control switch permits the pre-selection of a count sequence or batching requirement, read-out by four long life neon indicator tubes gives a clear and brilliant display. The counter can be supplied without neon tube indication but with the pre-selection feature, showing a suitable reduction in price.

Special features of this instrument, permitting different modes of operation are:

(1) Automatic re-set to zero immediately the pre-selected number is reached. The output relay may be held closed for an adjustable period between 20msec and 200msec.
(2) For single batching with manual re-set.
(3) Display time is adjustable between 20msec and 2sec when switched to automatic re-set.

The modular plug-in circuits allow for the choice of complete counting or batching and control systems to be constructed to a customer's specifications. This flexible system also allows for a fifth digit display and pre-selection switch to be added allowing a maximum count of 99999 without increasing the size of the case.

An internal Schmitt trigger ensures the counter working over a wide range of input waveforms. Optional frequency standards giving 4, 5, 6, 8 or 10c/s can be supplied.

EE 78 767 *for further details*

SELF-TUNING ULTRASONIC GENERATOR

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated below)

A self-tuning ultrasonic generator, the type 1191B Soniclean Automatic, is now available from Dawe Instruments Ltd. Designed for use with any one of a number of cleaning tanks such as the Dawe type 1165/126/8 (a five-gallon model with large radius corners having eight transducers and an effective cleaning area of 126in^2) the new generator comprises a transistor oscillator with transformer-coupled push-pull output stage. The output stage is in turn transformer-coupled to the transducer load, which forms an integral part of the standard tanks. A feedback circuit adjusts the input voltage of the amplifier according to the current in the motional arms of the transducers, causing the generator to seek and hold continuously the resonant frequency of the transducers, whatever the tank loading, temperature and other operating conditions. The only control necessary therefore is a simple on-off switch and warning light.

Most current ultrasonic cleaning systems are manually tuned to the most efficient operating point, which is the resonant frequency of the transducers. While this arrangement is perfectly satisfactory for small-scale cleaning operations, it is too slow for many production situations and requires a certain amount of experience. The self-tuning system of the type 1191B, on the other hand, automatically ensures maximum efficiency of operation at all times.

Because of the strains imposed by the permanent adjustment to resonance, a new type of transducer is used which employs lead zirconate titanate active elements instead of the more usual barium titanate. These type 1166 transducers are supplied already fitted to the standard tanks, but are also available separately. They have a particularly high conversion efficiency and can operate at temperatures as high as 300°F—more than sufficient for all foreseeable cleaning applications. Heated versions of the standard tanks are also available, with thermostatic control.

Pulses of ultrasonic energy at the frequency of the system (about 25kc/s) are produced by the type 1191B at a repetition rate of 100p/s or twice mains frequency, obtained by full-wave rectification. Peak power output is 600W, with about 300W average, supplied from a mains input of 600W. The generator is fully transistorized, and supplied in a bench-mounting metal cabinet measuring 16 × 14 × 6½in high. Weight is 40 lb.

EE 78 768 *for further details*

HIGH Q TUNING VARACTORS

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

The new ranges of high Q silicon tuning varactors have recently been introduced by Microwave Associates Ltd.

A series of alloyed epitaxial tuning varactors offer a minimum Q value of 100 at -4V, 50Mc/s, with a typical capacitance ratio of 5·2:1 for a bias change of 2 to 100V d.c. Capacitance values (defined at -4V) range from 6·5 to 47pF and peak reverse voltage is 110V minimum.

Q values of up to 350 (at -4V, 50Mc/s) are available in a wide range of diffused junction tuning varactors. Peak reverse voltage is 35V with a capacitance ratio of 3:1 between 2 and 30V. Higher voltage diodes can be supplied on request.

The subminiature glass encapsulation (wire-ended) is employed for both ranges and the micropill (0·055in height) is also offered for capacitance values of 15pF and lower. Maximum operating temperature is 150°C with a power dissipation of 500mW.

EE 78 769 *for further details*

PLUG-IN RELAYS

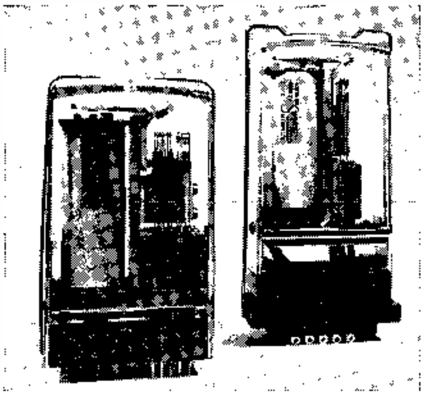
Simmonds Relays Ltd, Edinburgh Place, Harlow, Essex

(Illustrated below)

Introduced by Simmonds Relays Ltd are two plug-in versions of their type 3 000 relay. They have been designed to meet most applications, where fully enclosed dust-proof relays of this type are required.

Both types are enclosed in a moulded transparent polycarbonate cover and in each case the relays are held in their sockets by stainless steel retaining clips.

One version, type 3000/SP uses the tag ends of specially designed stub blades as the connecting medium. A moulded



phenolic socket is provided into which the relay plugs in. This relay is designed to fit the standard Post Office mounting ranks.

The other, type 3000/20P has the contact blades and coil terminals wired internally to a rectangular moulded 20-pin header plug which is securely fixed to the relay. Here again the necessary socket is provided with each relay.

A maximum of four heavy duty or eight light duty actions can be used with either relay. The contact ratings being the same for a standard type 3000 relay.

Special relays, housed in either of the above cases, are available. The following are a few examples from this category; Relays fitted with microswitches and open contacts; relays having contact insulation up to 10^{13}MΩ; slugged relays; remanent relays; relays with nickel iron sleeved cores to give high impedance to a.c.; relays with double, treble or sandwich wound coils.

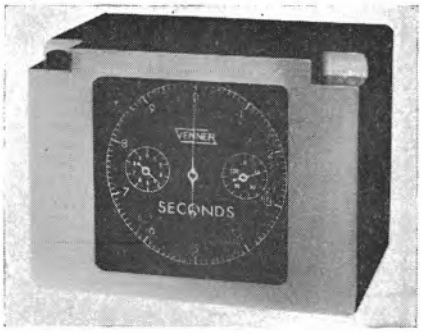
EE 78 770 *for further details*

ELECTRICAL STOP CLOCK

Venner Ltd, Kingston By-Pass, New Malden, Surrey

(Illustrated on page 129)

An electrical stop clock introduced by Venner Ltd provides a new and accurate means of direct experiment for physics



classes down to elementary level. It does away with the complicated processing of data necessary, for example, in all experimental methods previously used to determine the acceleration due to gravity or to study accelerated motion by use of a trolley.

The stop clock, based on an original design by J. N. Emery, Senior Physics Master of Trinity College, Glenalmond, is simple in operation and has electrical start/stop and electrical or manual re-set. It is synchronous motor-driven, operating from 12V 50c/s a.c. supply. Contacts provide an output pulse every second, and a continuous outgoing supply of 12V d.c. can also be obtained.

The main dial, with a diameter of 2½in. has white lettering on black and is calibrated in 1/100th seconds with a large sweep hand traversing once each second. Two small hands, one revolving once in 10sec and the other once in 150sec traverse inner dials; the 10sec dial is calibrated in 1sec divisions and the 150sec dial in 10sec divisions.

The stop clock is as accurate as the supply frequency.

All socket outlets and the manual reset push-button are set below the top level of the shock-resisting case for ease of stacking in bulk.

EE 78 771 *for further details*

COAXIAL ADAPTORS

Greenpar Engineering Ltd, Station Works, Cambridge Road, Harlow, Essex

(Illustrated below)

This new range of inter-series adaptors is available in any combination of series 'BNC', 'C', 'N' and 'UHF' plug and socket connexions. They can be supplied either in kit form, complete with two fixing spanners and fitted box, for laboratory or field use, or as sealed versions for standard applications. The kit offers the facility of making up any

four complete adaptors from a possible combination of 28 different types.

A NATO catalogue number has been allocated to each item in the range and they are fully interchangeable with any equivalent inter-series adaptors made to U.S. military drawings.

The v.s.w.r. performance has been considerably improved and with the exception of the series 'UHF' plug and socket combinations (which are not matched) the v.s.w.r. performance is better than 1·1 at 4 000Mc/s.

All inner contacts are gold plated.

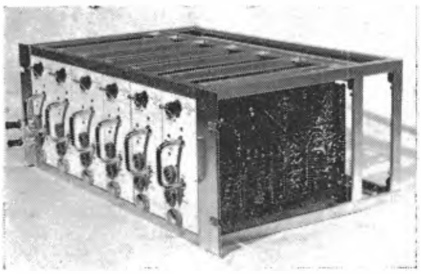
EE 78 772 *for further details*

SIX CHANNEL AMPLIFIER

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Illustrated below)

The new six channel amplifier system, type R6, is intended for the multi-channel amplification of low level signals from thermo-couples, strain gauges and transducers. The six amplifiers are contained in a 19in rack assembly. The mains connexions are made at the rear with the



input and output available on the front panels. The gains of the amplifiers are individually variable from 330 to 2 400. The amplifiers make use of photo resistive choppers giving the very low drift of less than $0{\cdot}2\mu V/°C$. The high input impedance of $100M\Omega$ prevents loading of the source, the high output current of 25mA and bandwidth of 20kc/s are suitable for driving ultra-violet recorders.

EE 78 773 *for further details*

V.H.F.-U.H.F. RECEIVER

VHF/UHF Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Illustrated above right)

The TR40/20 v.h.f.-u.h.f. receiver is a portable and fully transistorized instrument for the reception of a.m. speech transmissions.

It is designed for single channel operation in the 430 to 470Mc/s u.h.f. band and in addition it will operate on a single channel in the 70 to 90Mc/s v.h.f. band. It is intended for narrow band 25kc/s channel spacing, a triple superhet circuit being employed when used at u.h.f. and a double superhet at v.h.f. three stages of crystal control being utilized.

A small loudspeaker is provided and an internal 8V mercury cell will give approximately 70 hours use. 18 transistors and 4 diodes result in an extremely low battery consumption.

Plug-in whip aerials or external 75Ω aerial systems can be used. The TR40/20

weighs 4 lb and is 8in high, 7in wide and 2in deep.

A low power fully transistorized transmitter type TT40 is being developed for use with this receiver. Also, a combined portable transmitter/receiver type TRT/40 which will provide simplex or duplex operation.

EE 78 774 *for further details*

PRESSURE TRANSDUCER

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

The Consolidated Electrodynamics Division of Bell & Howell Ltd, has introduced the type 4-390 high output pressure transducer designed to perform reliably under extreme conditions of acceleration, vibration, and shock.

The 4-390 comprises an unbonded strain gauge sensing element with integral solid-state amplifier and power supply and provides a 5V d.c. output signal. It is designed for absolute and gauge measurements of fluids and gases compatible with 416 stainless steel, and is available in pressure ranges from $100lb/in^2$ up to $5\,000lb/in^2$.

The unit has no wipers or other mechanical parts to wear under vibration, and its power supply allows considerable fluctuation of the input power without degradation of the data signal.

Other design features include low-pass filtering to suppress high frequency noise in the output, a high degree of isolation between power in and signal out, and provision for prohibiting output voltages from exceeding a pre-set level.

The instrument is fully protected from accidental reversal of input power polarity, and electrical components are temperature compensated as an integral unit, thereby minimizing temperature effects.

Rated electrical excitation 28V d.c., 50mA maximum drain. Linearity and hysteresis are ±0·5 per cent of full range output, output impedance is less than 250Ω, and weight is 10 oz.

EE 78 775 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Science and General Division

Date: 2 February
Lecture: The Refurbishing of Engineers in the U.S.A.
By: J. K. Wolfe
Date: 11 February
Discussion: The Electrometer Amplifier: its Design and Applications
Opened by: G. I. Hitchcox and G. R. Taylor
Date: 16 February
Discussion: Universities under fire
Opened by: W. E. J. Farvis
Date: 19 February Time: 2.30 and 5 p.m.
Colloquium: Automatic Aids to Fault Finding in Computers
Date: 23 February
Discussion: What is Systems Engineering?
Opened by: K. L. Smith

Power Division

Date: 3 February
Lecture: A Seasonal Tariff for Domestic Supplies
By: D. J. Bolton
Date: 9 February
Discussion: The Effect of Variations in Supply Voltage in Industrial Processes
Opened by: J. A. Tatchell and H. J Sheppard
Date: 24 February
Lecture: Parallel Operation of Transformers with On-Load Tap changes and Negative-Reactance-Compounding Control
By: J. M. Cowan, K. L. Edmundson and L. L. Preston

Electronics Division

Date: 8 to 9 February
Conference on: Electronics Design
(All wishing to attend must register; forms available on application)
Date: 10 February
Lecture: Radar-Present Position and Future Trends
By: E. V. D. Glazier
Date: 15 February Time: 3 p.m.
Colloquium: Design of Solid State Power Supplies
(All wishing to attend must register; forms available on application)
Date: 18 February Time: 2.30 p.m.
Colloquium: Automatic Aids to Fault Finding in Computers
(All wishing to attend must register; forms available on application)
Date: 22 February
Lecture: Layer Structure of the Troposphere
By: J. A. Saxton, J. A. Lane, R. W. Meadows and P. A. Matthews

Faraday Lecture

Date: 17 February
Held at: Central Hall, Westminster, S.W.1
(Admission by ticket available from the I.E.E. on receipt of a stamped addressed envelope)
Lecture: Colour Television
By: F. C. McLean

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meetings will be held in the Institution's Lecture Room at 9 Bedford Square, London, W.C.1, and tickets will be required, unless otherwise stated.

Radar Group

Date: 17 February Time: 6 p.m.
Lecture: Transistorized Equipment Designed for Television Exploitation of Radar Information
By: R. Aste

Joint I.E.R.E.-I.E.E. Medical Electronics Group

Date: 10 February Time: 6 p.m.
Discussion: Safety of Operating Theatre Equipment

Joint I.E.R.E.-I.E.E. Computer Groups

Date: 18 February Time: 2.30 and 5.30 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2
Colloquium: Automatic Aids to Machine Fault Finding
(Registration will be necessary)

Television Group

Date: 24 February Time: 6 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1
Lecture: A Low Cost Video Tape Recorder for Professional Applications

North Western Section

Date: 11 February Time: 7 p.m.
Held at: Bolton Technical College, Bolton
Paper on: Analogue Computing and Digital Techniques in Aviation Control Systems

Southern Section

Date: 3 February Time: 7.30 p.m.
Held at: Bournemouth Municipal College of Technology and Commerce, Bournemouth
Lecture: Digital Computers in Road Traffic Control
By: H. A. Codd
Date: 18 February
Held at: Farnborough Technical College, Farnborough, Hampshire
Lecture: Hybrid Data Processing Systems at A.W.R.E.
By: E. W. Walker
Date: 23 February Time: 6.30 p.m.
Held at: University of Southampton
Lecture: Space Charge Limited Solid State Devices
(Joint meeting with the Local Centre of the Institution of Electrical Engineers)
By: G. T. Wright

South Western Section

Date: 24 February Time: 7 p.m.
Held at: University of Bristol Engineering Laboratories, Bristol
Lecture: Microminiaturization
(Joint meeting with the Bristol Branch of the Royal Aeronautical Society)
By: R. O. R. Chisholm
Date: 18 February Time: 7 p.m.
Held at: South Devon Technical College, Torquay
Lecture: Colour Television
(Joint meeting with the Western Centre of the Institution of Electrical Engineers)
By: H. V. Sims

South Wales Section

Date: 10 February Time: 6.30 p.m.
Held at: Welsh College of Advanced Technology, Cardiff
Lecture: Superconductors in Instrumentation
By: D. H. Parkinson

Scottish Section

Date: 10 February Time: 7 p.m.
Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh
Paper on: Sonar
By: V. G. Welsby
Date: 11 February Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Paper on: Sonar
By: V. G. Welsby

South Midland Section

Date: 26 February Time: 7 p.m.
Held at: B.B.C. Club, High Street, Evesham
Lecture: Design Aspects of Laboratory Test Instruments
By: K. A. Fletcher

East Midland Section

Date: 24 February Time: 6.30 p.m.
Held at: University of Leicester
Lecture: Parametric Amplifier
By: T. Buckley

Merseyside Section

Date: 17 February Time: 7.30 p.m.
Held at: Walker Art Gallery, Liverpool
Lecture: Airborne Navigational Aids
By: T. Gray

North Eastern Section

Date: 10 February Time: 6 p.m.
Held at: Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle
Lecture: Electronic Control Systems in Coal Mining
By: G. M. Rendall

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 2 February Time: 5 p.m.
Held at: The Conference Room, Fleet Building, 70 Farringdon Street, E.C.4.
Lecture: Some Problems of S.T.D. Working in Long Distance Area
By: F. C. Gould-Bacon

THE TELEVISION SOCIETY

Meetings are held at the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3. and commence at 7 p.m.
Date: 5 February
Lecture: The N.T.S.C. Colour Television System using Additional Reference Transmission
By: N. Mayer
Date: 18 February
Discussion: Six Months Experience with B.B.C.-2

PUBLICATIONS RECEIVED

SEMICONDUCTOR TEST WAVEFORM GENERATORS is a leaflet produced by EMI Electronics Ltd and which describes a range of equipment designed for the testing of cameras, monitors and other broadcast equipment. Operating on 405, 525 or 625 lines, this modular system comprises a mounting frame and combination of plug-in units. All waveform outputs are independent from one another and setting up controls are immediately accessible on the front panel of each unit. Use of semiconductors throughout ensures very low power consumption. Copies of this leaflet can be obtained from EMI Electronics Ltd, Hayes, Middlesex.

FORCED AIR COOLED U.H.F. POWER TETRODES is an STC application report dealing with the series 4X150/250 valves. There are four valves in this series and the characteristics for general use are described in this report. Copies are available from Standard Telephones and Cables Ltd, Special Valve Sales Department, Brixham Road, Paignton, Devon.

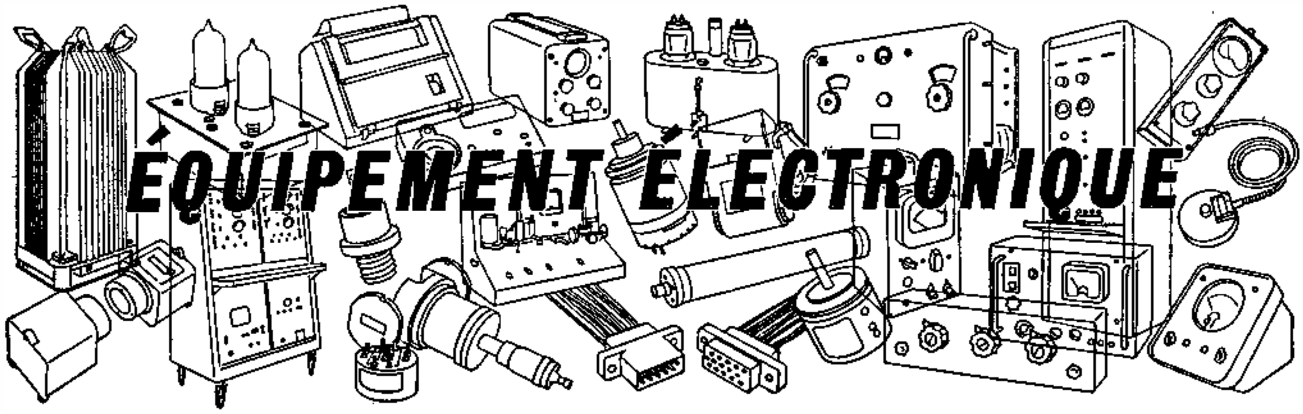
FILTERCONS—High Frequency Low Pass Filters—is a leaflet giving details and setting out the significant improvements that have been made in the ratings of some of the Filtercon range of high frequency low pass filters and the 1200 range of radio interference suppression filters. In most cases the low frequency current ratings have been doubled and in some cases they have been increased by a factor of five. Copies of this leaflet can be obtained from Erie Resistor Ltd, South Denes, Great Yarmouth, Norfolk.

RADIO FREQUENCY CABLES is a new publication from BICC. It is divided into two parts: the first part dealing with coaxial cables and the second part with BICC-BURNDY connectors and terminations. Copies of this publication are available from British Insulated Callenders Cables Ltd, 21 Bloomsbury Street, London, W.C.1.

LEAD ACID BATTERIES—Miscellaneous Uses—is an LDA publication which surveys the uses of these batteries in the U.K. outside the motor industry and the main stationary battery fields. Copies of this survey can be obtained from the Lead Development Association, 34 Berkeley Square, London, W.1.

A PROGRAMMED BOOK ON SEMICONDUCTOR DEVICES is a new linear programmed instruction book on semiconductor devices which has been published by the Mullard Educational Service. The book, believed to be the first of its kind published by an industrial organization, treats the subject in a non-mathematical way, emphasis being given to the physical aspects of electron and hole conduction. A supplementary booklet containing coloured diagrams illustrating semiconductor action is contained at the back of the book. Copies of the book (price 3s. 6d. including supplement) are available from the Mullard Educational Service, Mullard House, Torrington Place, London, W.C.1.

ÉQUIPEMENT ÉLECTRONIQUE



Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 124 à 129*

ALIMENTATION STABILISÉE

Kingshill Electronic Products Ltd, Torrens Works, Torrens Street, London, E.C.1

(Illustration à la page 124)

Un bloc d'alimentation stabilisée à transistors d'un modèle compact et à l'usage des laboratoires est fourni maintenant par la Kingshill Electronic Products Ltd. Sa sortie est de 0 à 50 V 1 A maximum et elle est à variation continue dans deux gammes à commutation. Le modèle 501 comprend des dispositifs de protection de source et de charge. Les circuits de protection entièrement électroniques peuvent être réglés de manière à limiter le courant de charge ou à couper la sortie complètement à huit valeurs de courant commutables de 5 mA à 1 A. Le choix de la fonction de limite ou de coupure s'effectue par commutateur à glisseur et la surcharge est indiquée par une lampe.

Le réglage du bloc est de 0,1 %, le rapport de stabilité de 1 000, l'ondulation et le bruit sont inférieurs à 1 mV et la température ambiante nominale est de 35° C au maximum. Chacune des sorties a deux bornes parallèles comportent des bases qui empêchent la dispersion. Les sorties, la tension et le courant peuvent être contrôlés avec précision sur l'indicateur à grand cadran de 0 à 25 V, de 0 à 50 V et de 0 à 1 A.

EE 78 751 *pour plus amples renseignements*

SPECTROMÈTRE DE RÉSONANCE QUADRIPOLAIRE NUCLÉAIRE

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

(Illustration à la page 124)

Le spectromètre de résonance quadripolaire nucléaire Scientifica a été réalisé spécifiquement pour répondre aux besoins des services scientifiques en ce qui concerne les appareils à résonance quadripolaire nucléaire pour les travaux de recherche et pour les démonstrations. Il peut être utilisé dans la gamme de fréquences de 10 à 60 MHz et détecter de nombreuses résonances quadripolaires nucléaires. En utilisant l'échantillon polycristallin standard fourni avec l'instrument, on peut obtenir de grands signaux de résonance à partir des deux isopopes de chlore Cl^{35} et Cl^{37} à des fréquences de 34, 233 et 27,007 MHz à une température de 25,2° C.

La mesure des fréquences r.q.n. dans des objets contenant plus d'un isotope donne une comparaison directe des moments quadripolaires nucléaires des différents isotopes et, pour certaines molécules, permet de déterminer leur valeur absolue. Pour les noyaux dont le moment quadripolaire est déjà connu, la mesure r.q.n. permet de prédire la répartition des électrons à l'intérieur de la molécule, ce qui fournit des indications sur la structure chimique de base de l'objet. On peut ainsi, par exemple, déterminer si un agglomérant chimique donné est purement covalent ou purement ionique. Pour un agglomérant de caractère mixte, on peut vérifier le pourcentage de contribution de chaque type.

Bien qu'aucun champ magnétique ne soit nécessaire en spectrométrie r.q.n. pure, on peut effectuer la séparation Zeeman de la ligne de résonance en appliquant un champ faible. La mesure de cette séparation pour différentes orientations de l'échantillon dans le champ magnétique permet de calculer la mesure dans laquelle le champ électrique moléculaire s'écarte de la symétrie axiale. Lorsqu'un champ magnétique hétérogène est appliqué à l'échantillon, le signal de résonance est étalé par l'élargissement de ligne qui en résulte. Cet effet peut être aisément montré en plaçant le petit aimant permanent fourni avec la trousse à environ 10 cm de l'échantillon.

Le processus général pour obtenir un signal de résonance quadripolaire nucléaire est comme suit: l'échantillon est placé à l'intérieur de la bobine d'un oscillateur à détection et la fréquence de ce dernier est continûment variée; si elle coïncide avec la valeur requise pour la résonance, l'échantillon absorbe l'énergie de l'oscillateur, provoquant ainsi un changement dans les conditions de fonc tionnement du circuit de l'oscillateur de détection. La fréquence est balayée de façon continue durant l'état de ré sonance, ce qui donne un signal d'ab sorption pouvant être amplifié pa l'amplificateur à transistors à faible brui incorporé au circuit. La sortie de ce dernier peut être injectée directement à un oscilloscope à faible gain afin d'y être affichée.

Pour pouvoir obtenir le rappor signal/bruit le plus élevé possible dans le spectromètre r.q.n. Scientifica, un circuit oscillateur de détection auto coupeur a été réalisé. Ce dernier com prend une commande pour modifier la fréquence de coupure, permettant ains de distinguer entre la fréquence de résonance centrale et ses bande latérales. Une autre commande, marqué "MOD. AMP." règle la largeur de band dans laquelle la fréquence d'oscillatio est balayée.

En plus de la sonde rigide d'échanti llon, utile à la mesure r.q.n. aux tem pératures de chambre, une sond flexible de 30 cm est également fournie Cette dernière permet de place l'échantillon dans un champ magnétiqu pour mesurer la séparation Zeeman Une longueur de fil argenté est égale ment fournie pour constituer des bobine supplémentaires répondant aux besoin individuels, soit pour recevoir de échantillons de différentes dimension soit pour le réglage approximatif de la fréquence de fonctionnement.

EE 78 752 *pour plus amples renseignement*

POTENTIOMÈTRE NUMÉRIQUE

Scientific Furnishings Ltd, Poynton, Cheshire

(Illustration à la page 124)

Le nouveau potentiomètre numériqu mis au point par la société Scientifi Furnishings Ltd allie l'équilibrage d zéro rapide et pratique par un seu bouton aux avantages de l'indicatio numérique de la tension mesurée. L décimale se déplace automatiquement au fur et à mesure que la gamme est changée d'un minimum de 0 à 9,99 mW à un maximum de 0 à 499,9 mV. Un adaptateur interchangeable étend la gamme jusqu'à 500 V avec des gammes de courant de 0 à 100 μA jusqu'à 5 mA. La sensibilité de mesure est supérieure à 0,2° C avec un thermocouple fer-constantan ou à 10 μV, avec une précision absolue de 0,15 %.

Une cellule standard intérieure Weston est fournie pour la normalisation rapide et une batterie au mercure de 14 Ah fournit une source de tension de précision pouvant atteindre 1 000 V pour l'étalonnage général et les essais. Un étui en cuivre du type caméra peut être fourni pour les mesures sur les lieux des travaux; pour ces dernières l'élément compact mesurant 13,97 cm × 14,60 cm × 19,68 cm est particulièrement indiqué. Le galvanomètre résistant au chocs est protégé contre les surintensités par le circuit à semi-conducteurs.

EE 78 753 *pour plus amples renseignements*

ALIMENTATION DE 400 Hz

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devonshire

(Illustration à la page 125)

Un nouveau bloc d'alimentation transistorisé à courant alternatif triphasé de 400 Hz a été réalisé par la société Hatfield Instruments Ltd. Il est destiné principalement à l'usage des fabricants de gyroscopes de précision. L'oscillateur de 400 Hz employé dans ce bloc est du type à multivibrateur avec une stabilité de ±1 % et il peut être verrouillé à une source extérieure si nécessaire. Le phasemètre fourni avec le bloc est largement indépendant de la tension. Cela facilite considérablement la précision du réglage jusqu'au déplacement de 120° entre les trois phases. La stabilité de la tension de sortie est de 0,2 %.

EE 78 754 *pour plus amples renseignements*

CONNECTEURS INDUSTRIELS

Plessey-UK Ltd, Ilford, Essex

(Illustration à la page 125)

La tendance générale vers l'automation, l'instrumentation et la télécommande dans l'industrie, particulièrement dans les industries lourdes telles que celles des mines et des pétroles, augmente les besoins en connecteurs électriques sûrs et robustes pouvant recevoir un certain nombre de circuits de signaux et de faible puissance.

Les gammes de connecteurs, coupleurs et pièces de montage de canalisation et de câble Plessey-UK, Mark II et IIF, ont été réalisées en liaison étroite avec le National Coal Board pour répondre à cette demande avec le maximum de sécurité et de résistance d'environnement, à un prix minimum. D'un encombrement des plus réduits, ils sont conformes à la spécification du National Coal Board pour les connecteurs multivoies et sont en cours d'adoption comme pièces standard. Ils sont actuellement utilisés dans les installations de mines R.O.L.F. et Collins ainsi que dans les systèmes téléphoniques et de transmission des mines du National Coal Board.

Le connecteur Mark 11 est basé sur le modèle bien connu des connecteurs Plessey UK-AN, dont des centaines de milliers d'exemplaires donnent pleine satisfaction dans des conditions d'emploi souvent ardues. La même qualité de contacts en laiton argentés ou dorés pour raccords de conducteurs soudés ou sertis, et les mêmes isolateurs en caoutchouc au silicone moulés sont utilisés dans des logements à brides boulonnées, de conception spéciale, faits de pièces de forge à haute résistance. Ce type de bride permet le montage sur panneau de fiches et de douilles ou de les utiliser comme pièces libres avec ou sans presse-étoupe de câble ou pièces de canalisation.

Les quatre modèles de base de connecteurs logent divers contacts d'une puissance nominale allant jusqu'à 12, 24, 50 et 100 A, de 2 à 48 directions. Des clefs de polarisation empêchent toute erreur d'accouplement, et l'accouplement croisé de connecteurs similaires est empêché par l'emploi de connecteurs orientés. Ces derniers comprennent six ouvertures filetées dans chaque bride à l'intérieur desquelles l'utilisateur peut visser trois broches d'orientation dans n'importe laquelle des vingt combinaisons prévues de manière à ce que les fiches ne s'accouplent qu'avec les douilles voulues.

Des presse-étoupe pouvant être vissés sur des fiches ou des douilles ont été spécialement conçus pour recevoir une gamme étendue de câbles blindés ou non blindés de différents diamètres dans chaque format de base. Ils sont serrés de façon étanche et indépendante sur l'enveloppe de câble extérieure et fixent le blindage à l'endroit voulu. Les pièces de canalisation fixent les canalisations souples et les conducteurs au connecteur.

La gamme de connecteurs et presse-étoupe Mark 11F (voir notre gravure) pour câbles blindés et non blindés constitue la version ignifuge du modèle Mark 11. Ils se conforment aux conditions exigées par la spécification britannique BS229 pour appareils électriques ignifuges et sont garanties par les certificats Buxton Nos. FLP 5344 et 5345 pour l'utilisation avec les gaz des groupes I, II et III.

EE 78 755 *pour plus amples renseignements*

INDICATEURS AUDIBLES

Highland Electronics Ltd, 26-28 Underwood Street, London, N.1

(Illustration à la page 125)

Les indicateurs audibles constitués de corps solides Sonalert à courant continu existent maintenant en une nouvelle version pour courant alternatif secteur de 200 à 250 V.

Ces dispositifs donnent un signal d'avertissement audible à haute fréquence, facilement identifiable et parfaitement distinct. Ils sont petits, de poids léger et autonomes, ne comportant qu'un seul trou de fixation. Ils ne contiennent pas de contacts mobiles, ne produisent pas de brouillages électriques, leur fiabilité est des plus élevées et ils sont prévus pour des conditions d'environnement difficiles. Ils mesurent 4,28 cm de diamètre × 5,71 cm de longueur.

Les indicateurs Sonalert ont une gamme étendue d'applications, comprenant la signalisation, le contrôle, l'alarme et l'avertissement.

EE 78 756 *pour plus amples renseignements*

SÉPARATEURS À RÉACTION

Moore, Reed & Co. Ltd., Woodman Works, Durnsford Road, London, S.W.19

(Illustration à la page 125)

La société Moore, Reed & Co. Ltd produit actuellement en série un nouveau séparateur compensé de 400 Hz, format 11, série 11RSF, pour le calcul de précision.

Ce séparateur dont la déviation à partir de l'onde sinusoïdale est inférieure à 0,05% et dont l'erreur entre axes est inférieure à 4 min, a pu réaliser en cours de production une proportion remarquablement élevée d'éléments à déviation inférieure à 0,03% et à niveau de bruit fondamental de 1 à 3 mV. Le rapport de transformation du stator et des enroulements de compensation est de 0,97.

Bien que d'une précision et d'"une qualité générale très élevées, ces séparateurs sont d'un prix réduit et sont fournis avec câbles mobiles ou avec plaquettes de connexions à 12 directions.

EE 78 757 *pour plus amples renseignements*

OSCILLATEUR À PHASES VARIABLES

Feedback Ltd, Crowborough, Sussex

(Illustration à la page 126)

L'oscillateur à phases variables VPO230 constitue un exemple d'application des méthodes d'amplification opérationnelle à un générateur à ondes sinusoïdales à gamme étendue.

Cet instrument se caractérise par ses trois sorties à réglage séparé: référence, quadrature et phase variable. Une fois réglées, ces sorties restent à une amplitude constante dans la gamme de fréquence de 1 Hz à 100 kHz. Les impédances de sortie sont de 6 Ω.

La sortie référence: quadrature est maintenue à 0,25° près de 90° et la phase variable peut être fixée à n'importe quel angle voulu à 1° près.

Lorsque l'appareil est utilisé comme oscillateur, les sorties ont une teneur en harmoniques inférieure à 0,5 %.

La technique de mesure de phase et d'amplitude comporte l'emploi d'un oscilloscope avec signal de phase variable appliqué au canal de déviation horizontal.

Le canal vertical est utilisé pour mesurer l'amplitude tandis que l'image Lissajous est réduite à 0 et que l'angle de phase est lu directement à partir du cadran de phases de l'oscillateur.

La souplesse d'emploi de l'appareil est énormément accrue par le dispositif de verrouillage extérieur. On peut ainsi asservir l'instrument à n'importe quelle autre source de fréquence dans sa gamme. Les mesures de phase sont donc possibles dans les systèmes fonctionnant à des niveaux de puissance élevés.

Lorsque l'instrument est utilisé dans des conditions d'asservissement, il se comporte comme un amplificateur sélecteur de fréquence et peut être accordé pour le choix de composants harmoniques à partir d'un signal principal complexe.

EE 78 758 *pour plus amples renseignements*

CHAMBRES DE CONTRÔLE D'ENVIRONNEMENT DE TEMPÉRATURE

Montyford Instruments Ltd, 79 Lots Road, London, S.W.10

(Illustration à la page 126)

Ces chambres d'essai on une gamme de —70° C à 200° et offrent un changement de température rapide pouvant être contrôlé à partir du panneau frontal. Un commutateur manuel automatique permet le changement de taux variable de température ainsi que le cyclage entre deux températures préréglées ou une stabilité élevée à une seule température.

Les programmes peuvent être arrêtés ou inversés à n'importe quel moment voulu. Le refroidissement est à base de CO_2 dont l'injection à l'échangeur de chaleur est réglée par une minuterie électrique. La chaleur est incorporée à la circulation du flux d'air comprimé, éliminant les points chauds et réduisant les gradients de température.

La chambre et le panneau de contrôle sont de construction modulaire et sont logés dans un bâti de largeur standard pouvant être agrandi pour pouvoir y monter des contrôleurs de programme, des amplificateurs ainsi que d'autres appareils pouvant être associés au matériel et aux composants soumis à l'essai.

On a prévu également la possibilité de connexions électriques à la chambre d'essai, la rendant ainsi propre au contrôle dynamique de composants et de matériel.

EE 78 759 *pour plus amples renseignements*

MESUREUR DE COUPLE

Mechatronics (London) Ltd, 317 Kennington Road, London, S.E.11

Un nouveau type de transducteur de couple a été réalisé par la société Mechatronics (London) Ltd.

Conçu en particulier pour le contrôle des moteurs F.H.P., des pompes et moteurs hydrauliques, des générateurs et des alternateurs, le transducteur consiste essentiellement en un torsiomètre qui rend superflu l'emploi du coussinet classique d'un dynamomètre à réaction.

Il fournit à la fois l'appui nécessaire au dispositif de puissance et un signal électrique proportionnel au couple de réaction développé par ce dispositif. Il comprend une plaque soutenue par une base de montage fixe au moyen de ressorts sensibles au couple. Des extensomètres électriques sont joints à ces ressorts, formant ainsi un pont extensométrique.

Les ressorts sont montés à angles de 45°. L'intersection d'une ligne tracée à travers les ressorts de flexion décrit le centre de rotation de la plateforme. Tout dispositif monté sur la plateforme et ayant son arbre d'entrée ou de sortie exactement au centre de rotation produira un couple de réaction qui sera enregistré par la plateforme.

Etant donné que l'appareil ne comporte ni coussinets ni bagues collectrices, la plateforme de couple n'a aucune perte par frottement ou limite de vitesse. Cette caractéristique permet à la plateforme de couple d'être utilisée à des couples très réduits sans erreur de frottement.

La résistance de pont est de 100 Ω et la tension de pont est de 12 V; la sortie nominale est de 2 mV/V. Les gammes de l'appareil vont de 0 à 3 livres/pouce à 0 à 300 livres/pouce.

EE 78 760 *pour plus amples renseignements*

SERVO COMPARATEUR NUMÉRIQUE

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

(Illustration à la page 126)

La fonction de ce comparateur est de comparer deux fréquences et de produire une sortie proportionnelle à leur différence. Cette sortie peut prendre la forme d'impulsions dont la fréquence représente la différence entre les fréquences originales ou elle peut être constituée par une tension analogue proportionnelle à la différence de fréquence.

Le dispositif règle d'abord les signaux d'entrée en amplitude et dans le temps puis soustrait une impulsion complète de la fréquence la plus élevée une fois par impulsion de la fréquence inférieure. Cette opération est indépendante du rapport de phase entre les deux signaux.

La forme d'onde de sortie est une série d'impulsions dont la largeur est de $\frac{1}{2}f_1$ et dont la fréquence est de $f_1 - f_2$. Cette forme d'onde est utilisée comme sortie numérique. Les impulsions sont alors réparties proportionnellement et égalisées. On obtient ainsi une sortie analogique.

L'avantage de ces méthodes réside dans le fait qu'elles ne comportent pas de circuit de minutage, la précision ne dépend donc que des temps de montée des dispositifs de sortie. On peut encore accroitre la souplesse d'emploi de l'appareil en ajoutant quelques éléments qui le rendent différentiel. Dans la première méthode, si $f_2 > f_1$, la sortie est de 0. Dans les autres méthodes la sortie est de $f_1 \sim f_2$ et l'on obtient également une indication du signal ayant la fréquence la plus élevée. L'affichage numérique peut être effectué simplement en dirigeant les sorties dans un registre à intervalles de temps fixe. En variante, un voltmètre peut être utilisé directement avec la sortie analogique.

Pour les applications telles que le contrôle automatique de vitesse, l'amplification de puissance peut être prévue jusqu'à 2,5 kW.

La fréquence fondamentale de l'appariel s'étend du courant continu à 10 MHz et les entrées peuvent varier entre 3 W et 100 W; l'impédance d'entrée est de 10 kΩ. La sortie est de 6 V à 1 mA (numérique) ou 0 à 6 W à 1 mA (analogique).

EE 78 761 *pour plus amples renseignements*

SONDE DE CONTRÔLE DE SEMICONDUCTEURS

Vacwell Engineering Co. Ltd, Sherman Road, Bromley North, Kent

(Illustration à la page 126)

La société Vacwell Engineering Co. Ltd vient d'ajouter une nouvelle pièce à sa gamme d'appareils pour l'industrie des semiconducteurs. Ce composant qui s'appelle sonde de laboratoire type PR.61 a été conçu essentiellement pour le contrôle-surprise de dés se trouvant encore sous forme de galette, mais il a également d'autres usages en particulier dans les micro-circuits.

L'instrument est monté sur bande d'essai et il est normalement fourni avec deux têtes de sonde bien qu'on puisse en fixer quatre. Les sondes sont isolées électriquement du corps principal et sont élevées ou baissées sur la galette en actionnant un petit levier manuel. Une sonde motorisée à part, avec commande à bouton poussoir et comprenant en outre un dispositif à réservoir d'encre permet de marquer sur place toute galette défectueuse.

Un mandrin à vide isolé, monté sur une plateforme X-Y réglable par micromètre, tient la galette soumise à l'examen et peut être mis en rotation pour faciliter l'alignement des galettes.

La spécification du PR.61 qui peut être facilement accouplé à un contrôleur de transistor Vacwell à trois paramètres et semi-automatique ou à n'importe quel autre appareil de contrôle, comprend un microscope binoculaire d'amplification de puissance 50 et l'éclairage de la plateforme de travail. L'accessoire à microscope peut évoluer à distance de la plateforme pendant que l'on place la galette.

EE 78 762 *pour plus amples renseignements*

COMMUTATEUR À LAME VIBRANTE SUB-MINIATURE

Flight Refuelling Ltd, Wimborne, Dorset

(Illustration à la page 127)

La société Flight Refuelling Ltd. dispose maintenant du dernier modèle "A" de commutateurs à lame vibrante de la société Hamlin. Ce nouveau commutateur à lame vibrante sub-miniature, le MTRG-2, fait le pendant du commutateur récemment introduit sous le nom de MRG-DT forme "C." Etant donné que les espaces de contact des deux aucune polarisation magnétique et vu que les espaces de contact des deux commutateurs sont alignés, on peut maintenant obtenir la permutation bipolaire, la permutation unipolaire ainsi que les combinaisons unipolaires à une direction dans un logement minimum.

La longueur hors-tout de ces nouveaux commutateurs est de 5,39 cm, le diamètre du verre étant de 2 mm et les conducteurs pouvant être réduits à 1,8 cm. Ces commutateurs on hérité toutes les caractéristiques d'environnement des commutateurs à lame vibrante de format standard et leur durée de vie à pleine charge est de 3 W c.c. résistif: 28 V maximum, 0,250 A maximum, de 10×10^6 opérations avec résistance de contact à 0,150 Ω maximum initial et 0,100 Ω nominal.

Pour permettre de comparer les dimensions, notre gravure montre le commutateur sub-miniature MTRG-D à côté du commutateur standard forme "A" DRG-2.

EE 78 763 *pour plus amples renseignements*

CONTRÔLEURS DE TEMPÉRATURE

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(*Illustration à la page 127*)

La société Kent Precision Electronics Ltd vient de produire une gamme d'instruments de contrôle de température constitués de corps solides d'un dessin avancé pour le contrôle de température de hauts fourneaux électriques jusqu'à 2 000° C avec une précision de 0,5 %. Le système se compose d'éléments transducteurs de température classique qui injectent un signal proportionnel à la température dans un amplificateur de courant continu à transistors de précision. Ce signal d'entrée est ensuite comparé à un signal de référence stable et précis, dont la grandeur est réglée selon les besoins sur une échelle indicatrice à l'avant de l'instrument. La tension d'erreur est utilisée pour contrôler le système de sortie de puissance à thyristors afin de régler la puissance livrée aux éléments de chauffage.

Pour les puissances inférieures, le contrôle proportionnel de la puissance de sortie est effectué en variant l'angle de phase auquel le circuit à thyristors est amorçé durant chaque demi-cycle de la tension secteur alternative. Pour les puissances élevées, et pour surmonter le brouillage, les pointes et les facteurs de faible puissance, le circuit à thyristors est commuté pour produire des demi-cycles intégraux, le contrôle proportionnel étant effectué en variant le rapport du nombre des cycles arrêt et marche. Dans tous les cas, le contrôle proportionnel est opérationnel sur 3 % environ de l'échelle.

Pour le contrôle de courant supérieur à 100 A, lorsque le contrôle par thyristors s'avère peu rentable, des circuits de sortie à thyristors de faible puissance de type standard peuvent être utilisés pour contrôler des éléments à réaction saturables pour le contrôle proportionnel ou, en variante, des contacteurs électromagnétiques pour le fonctionnement arrêt-marche. Notre gravure montre un contrôleur de température proportionnel standard de 10 kW.

EE 78 764 *pour plus amples renseignements*

POTENTIOMÈTRE PRÉRÉGLÉ MINIATURE

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

Le potentiomètre type 62 est un nouvel instrument préréglé et ajustable par tournevis réalisé par Morganite Resistors Ltd. Il a été spécialement conçu pour les équipements miniature dans lesquels les composants de dimensions réduites sont de prime importance. Le potentiomètre type 62 mesure environ 12 × 9 mm.

L'espacement des cosses est prévu pour percer des ouvertures modulaires de 2,5 mm. Le contact peut être réglé des deux côtés du composant et la piste de résistance est protégée contre l'endommagement accidentel durant le réglage.

Les deux modèles présentés peuvent être fixés directement sur la plaquette de circuit imprimé. Le type 62H est fixé en position verticale.

La puissance nominale est de 0,2 W à 40° C de température ambiante et la gamme de résistance est de 100 Ω à 1 MΩ.

EE 78 7.5 *pour plus amples renseignements*

TRANSISTORS DE COUPLE

Saunders-Roe Division, Westland Aircraft Ltd, Yeovil, Somerset

(*Illustration à la page 127*)

Les nouveaux transducteurs de couple mark 4, type 2, comprennent des assemblages de stators et de rotors construits en une pièce de fibre de verre moulée. Le couple est transmis à travers un arbre d'acier à haute résistance sur lequel sont montés des extensomètres à feuille Saunders-Roe. La connexion électrique entre le rotor et le stator s'effectue par collecteurs d'argent et par balais de graphite d'argent, puis par câble à quatre noyaux à l'appareil d'instrumentation. Les erreurs totales du transducteur dues à la linéarité, à l'hystérésis et à la répétabilité sont inférieures à ±0,2 %.

L'instrumentation de ce transducteur est fournie par l'indicateur TM6 qui produit deux gammes de couple d'échelle totale à partir de n'importe quel transducteur. En dehors de la lecture directe Cirscale du couple moyen effectif, des douilles sont prévues pour la connexion à des enregistreurs auxiliaires, à des oscilloscopes et à d'autres appareils supplémentaires facultatifs. La précision de lecture du couple moyen effectif n'est pas affectée par les grandes variations cycliques du couple.

Le système fonctionne à partir de la vitesse nulle jusqu'à la vitesse nominale maxima selon la grandeur du transducteur et il a une réponse de fréquence très élevée pour les recherches cycliques et transitoires. L'indicateur peut être actionné soit par courant alternatif secteur soit par batterie. Les transducteurs sont fournis pour 17 puissances nominales variant de 0 à 5 livres par pied à 0 à 50 000 livres par pied, et pour des vitesses de 0 jusqu'à 8 000 tours/min à 0 jusqu'à 800 tours/min.

EE 78 766 *pour plus amples renseignements*

COMPTEUR DE LOTS

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(*Illustration à la page 128*)

Cet instrument portatif et compact permet le comptage de lots jusqu'à 100 kHz. C'est un appareil à plaquettes de circuit imprimé et à système à fiches. Il est entièrement muni de dispositifs au silicium.

Son commutateur de contrôle par la bande permet la présélection d'une séquence de comptage, et la lecture au moyen de quatre tubes indicateurs au néon de longue durée assure un affichage net et distinct. Il peut être fourni sans tube indicateur au néon mais avec dispositif de pré-sélection et à prix réduit en conséquence.

Les caractéristiques spéciales de ce compteur qui permettent différents modes de fonctionnement sont les suivantes:

(1) Réenclenchement automatique à zéro dès que le nombre présélectionné a été atteint. Le relais de sortie peut être maintenu fermé pour une période réglable entre 20 msec et 200 msec.

(2) Pour contrôle par l'eau simple avec réenclenchement manuel.

(3) Le temps d'affichage est réglable entre 20 msec et 2 sec lorsque l'appareil est commuté au réenclenchement automatique.

Les circuits à fiches modulaires permettent le choix du comptage complet et la construction de systèmes de contrôle selon les spécifications du client. Ce système d'une grande souplesse permet également un affichage à cinquième chiffre et l'adjonction d'un commutateur de pré-sélection permettant un comptage maximum de 99999 sans augmenter les dimensions du coffret.

Un déclencher intérieur Schmitt assure le fonctionnement du compteur dans une gamme étendue de formes d'ondes d'entrée. Un étalon de fréquence facultatif donnant 4, 5, 6, 8 ou 10 Hz peut être fourni sur demande.

EE 78 767 *pour plus amples renseignements*

GÉNÉRATEUR À ULTRASONS D'AUTO-ACCORD

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(*Illustration à la page 128*)

La société Dawe Instruments Ltd fournit maintenant un générateur à ultrasons d'auto-accord, le modèle 1191 B Soniclean Automatic. Conçu pour l'emploi avec n'importe quel nombre de réservoirs de dégraissage tels que le réservoir Dawe type 1165/126/8 (modèle à 5 gallons avec coins à grand rayon comprenant 8 transducteurs et une surface de dégraissage effective de 126 pouces carrés), le nouveau générateur comprend un oscillateur à transistors avec étage de sortie push-pull accouplé au transformateur. L'étage de sortie est à son tour accouplé par transformateur à la charge du transducteur qui forme une partie intégrante des réservoirs standard. Un circuit à réaction régle la tension de l'amplificateur selon le courant dans les bras à mouvement des transducteurs, ce qui amène le générateur à chercher et à maintenir de façon continue la fréquence résonante des transducteurs, quelle que soit la charge du réservoir, la température et les autres conditions de fonctionnement. La seule commande nécessaire est donc un commutateur arrêt-marche et un voyant lumineux.

La plupart des systèmes de dégraissage par ultrasons du modèle courant sont accordés manuellement au point de fonctionnement le plus efficace, constitué par la fréquence de résonance des transducteurs. Bien que ce montage soit parfaitement adéquat pour les opérations de dégraissage sur échelle réduite, il est néanmoins beaucoup trop lent pour la plupart des applications de production et exige une certaine expérience. Le système d'autoaccord du type 1191B, d'autre part, assure une efficacité maxima automatique du fonctionnement à n'importe quel moment.

En raison des contraintes imposées par le réglage permanent à la résonance, un nouveau type de transducteur a été utilisé. Ce dernier emploie des éléments actifs au titanate de zirconium de plomb à la place du titanate de barium habituel. Ces transducteurs type 1166 sont fournis déjà montés sur les réservoirs standard mais ils peuvent être également fournis séparément. Leur efficacité de conversion est particulièrement élevée et ils peuvent être utilisés à des températures atteignant 300° F c'est à dire à une température plus que suffisante pour la plupart des opérations de dégraissage habituellement prévues. Il existe en outre des versions chauffées des réservoirs standard, avec contrôle thermostatique.

Les impulsions d'énergie à ultrasons à la fréquence de résonance du système (environ 25 kHz) sont produites par le type 1191B à une vitesse de répétition de 100 p/sec ou deux fois la fréquence secteur, obtenue par redressement sur la totalité de l'onde. La sortie de puissance de pointe est de 600 W, avec environ 300 W de moyenne fourni par une entrée secteur de 600 W. Le générateur est entièrement transistorisé et fourni dans un coffret métallique pour montage sur banc d'essai mesurant 40 cm × 35,5 cm × 16,5 cm de hauteur. Son poids est d'environ 18 kg.

***EE 78 768** pour plus amples renseignements*

VARACTORS D'ACCORD À Q ÉLEVÉ

Microwave Associates Ltd, Cradock Road Luton, Bedfordshire

Deux nouvelles gammes de varactors d'accord au silicium à Q élevé viennent d'être introduites par la société Microwave Associates Ltd.

La série A de Varactors d'accord épitaxiaux en alliage offre une valeur Q minima de 100 à −4 V, 50 MHz, avec un rapport de capacité typique de 5,2 : 1 pour un changement de polarisation de 2 à 100 V c.c. Les valeurs de capacité (définies à −4 V) s'étendent de 6,5 à 47 pF et la tension d'inversion de pointe est de 110 V au minimum.

Des valeurs Q pouvant atteindre 350 (a −4 V, 50 MHz) sont fournies dans une gamme étendue de Varactors d'accord à jonction par diffusion. La tension d'inversion de pointe est de 35 V avec un rapport de capacité de 3 : 1 entre 2 et 30 V. Des diodes pour tensions plus élevées peuvent être fournies sur demande.

L'encapsulage sous verre sub-miniature (à extrémité de fil métallique) est employé pour les deux gammes et un élément microminiature est également offert pour des valeurs de capacité de 15 pF et pour des valeurs inférieures à ce chiffre. La température de fonctionnement maxima est de 150° C avec une dissipation de puissance de 500 mW.

***EE 78 769** pour plus amples renseignements*

RELAIS INTERCHANGEABLES

Simmonds Relays Ltd, Edinburgh Place, Harlow, Essex

(Illustration à la page 128)

Il s'agit ici de deux versions interchangeables du relais type 3000 réalisé par Simmonds Relays Lt. Ces relais à fiches répondent à la plupart des applications exigeant des éléments anti-poussière et entièrement enfermés.

Les deux types sont enfermés dans un boîtier moulé au polycarbonate transparent et les relais sont maintenus dans leur douille au moyen de pinces en acier inoxydable.

Le type 300/SP comporte des extrémités à broches de contact constituées par des lames spécialement conçues pour servir d'éléments de connexion. Une douille moulée phénolique est fournie pour pouvoir enficher le relais; ce dernier a été conçu pour pouvoir être adapté aux bâtis de montage standard de l'administration des postes.

Quand au relais type 3000/20P, ses lames de contact et ses bornes de bobines sont bobinées intérieurement sur une fiche rectangulaire moulée à 20 broches fermement fixée au relais. Ici encore la douille nécessaire est fournie avec chaque relais.

Un maximum de quatre opérations de puissance ou de huit opérations légères peuvent être effectuées avec chacun des relais, les valeurs nominales de contact étant les mêmes que pour le relais standard type 3000.

Il existe également des relais spéciaux, logés dans l'un ou l'autre des deux coffrets. Nous donnons ci-après quelques exemples de cette catégorie: relais muni de micro-rupteurs et de contacts ouverts; relais à isolement de contact jusqu'à 10^{13} MΩ; relais à cartouche; relais à persistance; relais à noyau à manchon de nickel fer donnant une impédance élevée au courant alternatif; relais avec bobine double, triple ou bobinée en sandwich.

***EE 78 770** pour plus amples renseignements*

CHRONOMÈTRE ÉLECTRIQUE

Venner Ltd, Kingston By-Pass, New Malden, Surrey

(Illustration à la page 129)

Le chronomètre électrique réalisé par la société Venner Ltd constitue un nouveau moyen de précision pour les expériences directes effectuées dans les classes de physique jusqu'au niveau élémentaire. Il dispense du traitement compliqué de données qui est nécessaire, par exemple dans toutes les méthodes expérimentales employées jusqu'ici pour déterminer l'accélération dûe à la gravité ou pour l'étude du mouvement accéléré au moyen d'un chariot.

Ce chronomètre qui est basé sur la conception originale de J. N. Emery, professeur de physique à Trinity College, Glenalmond, est d'un fonctionnement fort simple et comporte l'arrêt-marche électrique et le réenclenchement manuel ou électrique. Il est actionné par moteur synchrone, fonctionnant sur courant alternatif de 12 V, 50 Hz. Les contacts fournissent une impulsion de sortie toutes les secondes et on peut aussi obtenir une alimentation de sortie continue de 12 V c.c.

Le cadran principal, dont le diamètre est de 6,35 cm, est à lettrage blanc sur fond noir et il est étalonné en 1/100e de seconde avec un grand index de balayage effectuant un tour chaque seconde. Deux petites aiguilles, dont l'une effectue un tour en dix secondes et l'autre en 150 secondes traverse les cadrans intérieurs; le cadran de 10 sec est étalonné en divisions de 1 seconde et le cadran de 150 sec est étalonné en divisions de 10 sec.

Le chronomètre est aussi précis que la fréquence d'alimentation.

Toutes les sorties de douilles ainsi que le bouton poussoir de réenclenchement manuel sont réglés au-dessous du niveau supérieur du boîtier résistant au choc afin de faciliter l'empilage en vrac.

***EE 78 771** pour plus amples renseignements*

ADAPTATEURS COAXIAUX

Greenpar Engineering Ltd, Station Works, Cambridge Road, Harlow, Essex

(Illustration à la page 129)

Cette nouvelle gamme d'adaptateurs inter-serie est livrable dans n'importe quelle combinaison des séries de connexions à fiche et douille "BNC", "C" "N" et "UHF". Ils peuvent être fournis soit sous forme de trousse, complète avec deux clefs de fixation et boîtier pour laboratoires ou travaux pratiques, ou comme versions scellées pour les applications courantes. La trousse offre l'avantage de pouvoir réaliser n'importe quels quatre adaptateurs complets à partir d'une combinaison de 28 types différents.

Un chiffre de catalogue OTAN a été donné à chacune des pièces de la gamme et ces dernières sont totalement interchangeables avec n'importe quelle série d'adaptateurs inter-série construits selon les dessins militaires américains.

Le comportement du rapport d'amplitude de tension a été considérablement amélioré et, à l'exception des combinaisons de fiches et douilles de la série "UHF" (qui ne sont pas adaptées), le comportement du rapport d'amplitude de tension est supérieur à 1,1 à 4000 MKz.

Tous les contacts intérieurs sont dorés.

EE 78 772 *pour plus amples renseignements*

AMPLIFICATEUR À SIX VOIES

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Illustration à la page 129)

Ce nouvel amplificateur à six voies, type R6, a été conçu pour l'amplification multivoies de signaux à faible niveau provenant de thermocouples, d'extensomètres et de transducteurs. Les six amplificateurs sont contenus dans un assemblage pour bâti de 48,25 cm. Les connexions au secteur se trouvent à l'arrière tandis que l'entrée et la sortie se trouvent sur les panneaux frontaux. Le gain des amplificateurs est à variations individuelles de 330 à 2400. Les amplificateurs comprennent des interrupteurs photorésistants donnant une dérive inférieure à 0,2 μV/° C. L'impédance d'entrée élevée de 100 MΩ empêche la charge de la source, tandis que le courant de sortie élevé de 25 mA et la largeur de bande de 20 kHz permettent d'actionner des enregistreurs à rayons ultraviolets.

EE 78 773 *pour plus amples renseignements*

RÉCEPTEUR V.H.F.-U.H.F.

VHF/UHF Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Illustration à la page 129)

Le récepteur v.h.f.-u.h.f. TR40/20 est un instrument portatif et entièrement transistorisé pour la réception des émissions de parole modulées en amplitude.

Il a été conçu pour le fonctionnement univoie dans la bande de 430 à 470 MHz UHF et il peut en outre fonctionner sur une seule voie dans la bande VHF de 70 à 90 MHz. Il a été prévu pour l'espacement de voies dans la bande étroite de 25 kHz, un triple circuit superhétérodyne étant utilisé lorsque l'appareil est employé sur UHF et une double superhétérodyne pour la très haute fréquence. Trois étages de contrôle piézoélectrique sont également utilisés.

L'appareil comprend un petit hautparleur ainsi qu'une petite cellule intérieure au mercure de 8 V assurant une utilisation d'environ 70 heures. 18 transistors et 4 diodes donnent une consommation de batterie extrêmement faible.

On peut employer soit des antennes à fiche soit des antennes extérieures de 75 Ω. Le récepteur TR40/20 pèse moins de 2 kg et mesure 20,32 cm de hauteur, 17,78 cm de largeur et 5,08 cm de profondeur.

Un émetteur transistorisé type TT40 à faible puissance a été mis au point pour l'emploi avec ce récepteur de même qu'un émetteur/récepteur combiné type TRT/40 qui permettra le fonctionnement en simplex ou en duplex.

EE 78 774 *pour plus amples renseignements*

TRANSDUCTEUR DE PRESSION

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

La consolidated Electrodynamics Division de la société Bell & Howell Ltd a réalisé un transducteur de pression à sortie élevée, type 4-390, conçu pour assurer un fonctionnement sûr dans des conditions extrêmes d'accélération, de vibration et de choc.

Le transducteur 4-390 comprend un élément sensible d'extensomètre avec amplificateur et alimentation constitués de corps solides faisant partie intégrante du transducteur et fournissant un signal de sortie de 5 V c.c. Il a été étudié pour la mesure absolue et de jauge de liquides et de gaz compatibles avec l'acier inoxydable 416 et il peut être livré pour des gammes de pression de 100 livres/pouces carrés jusqu'à 5000 livres/pouces carrés.

Il ne comprend ni frotteur ni autre partie mécanique sujette à l'usure par vibration, et son alimentation tolère des fluctuations considérables de la puissance d'entrée sans pour cela entraîner une dégradation du signal d'entrée.

Sa conception se caractérise en outre par un filtrage passebas supprimant le bruit de haute fréquence à la sortie, un degré élevé d'isolement entre la puissance d'entrée et le signal de sortie et la possibilité d'empêcher les tensions de sortie de dépasser un niveau préréglé.

Le transducteur 4-390 est entièrement protégé contre l'inversion accidentelle de la polarité de la puissance d'entrée et les composants électriques sont à compensation de température, réduisant ainsi les effets de température.

L'excitation électrique nominale est de 28 V c.c., 50 mA de consommation maxima. La linéarité et l'hystérésis sont de $\pm$ 0,5 % de la sortie sur la totalité de la gamme. L'impédance de sortie est inférieure à 250 Ω et le poids du transducteur est d'environ 311 grammes.

EE 78 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Machine à effet d'évanouissement pour la simulation de l'ionosphère par D. Beck et J. A. Betts

Résumé de l'article aux pages 74 à 79

Cette machine à effet d'évanouissement simule les variations aléatoires d'amplitude et d'angle de phase d'un signal de radio HF à longue distance en ajoutant les sorties de huit oscillateurs à cristal nominalement à 100kHz. Compte est tenu de la propagation simultanée sur deux voies avec une différence de temps de voie allant jusqu'à 2 millisecondes. Dans sa forme actuelle, cette machine permet l'emploi d'une seule antenne mais un dédoublement de l'équipement pourrait assurer une réception à double diversité. Le bruit blanc de Gauss peut être ajouté au signal afin de fournir des rapports de puissance signal-bruit de 0–50dB. L'entrée, dans la gamme de 300–6 000Hz peut être soit un radiotélégraphe soit n'importe quel signal de bande latérale unique.

Le transistor à effet de champ utilisé comme interrupteur à faible niveau par K. Barton

Résumé de l'article aux pages 80 à 83

L'interrupteur de transistors à effet de champ est comparé avec l'interrupteur de transistors à alliage de silicium classique. En utilisant un interrupteur de transistor à effet de champ, il est démontré qu'une dérive de tension inférieure à 0,1μV/°C, et une dérive à long terme inférieure à 0,1μV par mois peut être réalisé facilement sans l'emploi de procédés de sélection coûteux ou d'un four à température contrôlée.

Les appendices comprennent les calculs de gain avant de l'interrupteur de transistors à effet de champ, pour des entrées de courant et de force électromagnétique, ainsi que le calcul de tension de flanquement dû au courant de flanquement.

Résistances fixes pour tensions élevées par P. L. Kirby

Résumé de l'article aux pages 84 à 89

L'auteur étudie les moyens par lesquels la géométrie d'une résistance à pellicule peut être contrôlée pour réduire les variations de tension. Il examine le problème de l'absence de linéarité et les limites d'un "coefficient de tension". Pour les applications à haute tension, il est pratique de combiner les effets de tension et de coefficient de température. L'auteur décrit ensuite un modèle perfectionné de résistance à haute tension basée sur une pellicule épaisse d'oxyde d'étain "dilué". Le niveau de bruit est remarquablement faible et les résistances sont stables jusqu'à ±1% environ dans des conditions normales et peuvent tolérer des impulsions de tension atteignant 50kV de pointe. Des éclateurs peuvent être fournis pour assurer la protection contre les surtensions. Le comportement de ces résistances dans la chaîne de fuite d'un générateur électrostatique à haute tension est décrit.

Le gain de puissance de réseaux d'antennes Yagi par P. Knight et R.D.C. Thoday

Résumé de l'article aux pages 89 à 90

Les réseaux d'antennes Yagi à grande ouverture sont fréquement utilisés dans les bandes VHF et UHF afin d'accroître le gain et la directivité. La méthode classique de calcul du gain basée sur l'intégration du système de radiation de puissance n'est pas facile à appliquer en raison des formes compliquées de ces réseaux. Cet article décrit une nouvelle méthode pour le calcul de gain basée sur l'estimation de la résistance réciproque entre paires de Yagi. Il contient également un tableau permettant de déterminer aisément les gains d'une gamme de réseaux Yagi.

Une étude sur les méthode d'accélération de post-déviation utilisées dans les tubes cathodiques d'instruments par K. J. Ming

Résumé de l'article aux pages 94 à 97

L'auteur donne un aperçu du fonctionnement des rubes cathodiques communément utilisés dans les instruments électroniques actuels. Ces tubes comportent généralement une forme quelconque d'accélération de post-déviation, soit du type à bande simple soit du type à spirale de graphite.

Des tubes ayant une sensibilité de déviation supérieure peuvent être obtenus actuellement à la suite de la réalisation de systèmes perfectionnés d'accélération de post-déviation utilisant un réseau ou une électrode de maille. L'auteur décrit les caractéristiques essentielles de trois de ces systèmes, à savoir l'accélérateur axial, l'accélérateur radial et l'accélérateur à lentille divergente.

Un convertisseur analogique/numérique graphique par R. L. J. Gilbert et N. S. Oakey

Résumé de l'article aux pages 98 à 101

Le matériel décrit dans cet article peut lire de manière automatique les enregistrements sur diagramme effectués par le procédé à style chaud, et présenter les renseignements sous forme numérique. La géométrie du système de balayage resemble à celle d'un analyseur à spot mobile et la position de la trace est mesurée par une minuterie-compteur. La sortie peut être enregistrée par n'importe quel enregistreur numérique comme par exemple un imprimeur à lignes. La précision est de ±0,5% sur la totalité de l'échelle et le taux de répétition est de ±0,1%.

Système de sélection de piste de cellule magasin à tambour magnétique par S. C. Acharyya

Résumé de l'article aux pages 102 à 105

Cet article traite de la construction d'un commutateur à sélection de piste pouvant être employé dans un système d'emmagasinage à tambour magnétique. L'emploi de déclencheurs à transistors dans le commutateur impose quelques restrictions au choix des valeurs de composants, en particulier lorsque deux niveaux de tension sont utilisés.

Avec les bobinages de tête à faible inductance, le commutateur s'avère simple, sûr et économique. Avec un tambour magnétique ayant une inductance de bobinage de tête plus élevée, la nécessité de devoir employer des transistors à haute tension rend le commutateur relativement plus coûteux, bien que la simplicité et la fiabilité du dispositif soit conservées.

Retard thermique amélioré par un nouveau principe de circuit par G. F. Klepp et J. E. Brooke-Stewart

Résumé de l'article aux pages 106 à 149

Les commutateurs à retard thermique trouvent des applications dans la connexion différée de H.T. à des tubes chauffés indirectement et à des redresseurs, ainsi que dans la mise en marche de moteurs, dans la commutation à répétition et dans de nombreux autres circuits à retard.

Les commutateurs à retard thermique ont deux caractéristiques principales: le temps de lancement et le temps d'ouverture. Si le retard final est réalisé non pas par le temps de lancement seulement mais en additionnant le temps de lancement et le temps d'ouverture, l'étalement de temps entre les commutateurs individuels sera alors réduit par rapport à l'une ou l'autre de ces deux caractéristiques. Il est démontré que le temps de retard est indépendant dans une grande mesure de la tension de chauffage et de la température ambiante. Il subit, en outre, moins de variations en fonction de la durée du mécanisme de commutation.

Un circuit cascode à paire complémentaire pour amplification H.F. par M. D. Wood

Résumé de l'article aux pages 110 à 111

Il est généralement reconnu qu'un circuit cascode à base à la masse/émetteur à la masse peut être utilisé en variante à la neutralisation. Cet article compare les avantages de la cascode et de la neutralisation et propose un circuit cascode à air complémentaire qui comporterait certains avantages par rapport au circuit cascode normal, particulièrement aux fréquences supérieures à 10-20MHz.

Un amplificateur à courant continu avec gain de tension unitaire par G. G. Bloodworth

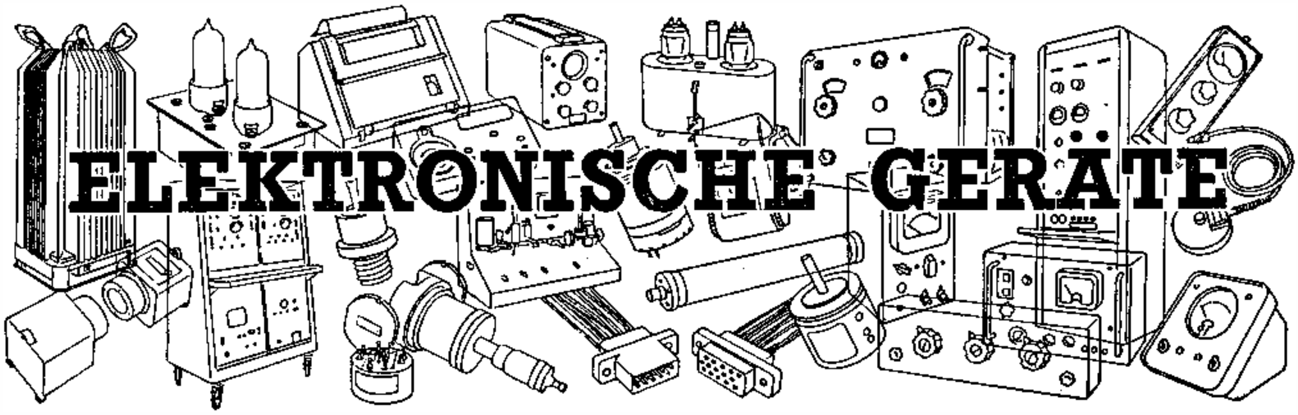
Résumé de l'article aux pages 112 à 114

On a mis au point un amplificateur intermédiaire à couplage direct dont le gain de tension est très proche de l'unité, sans flanquement entre l'entrée et la sortie. On peut en outre annuler la dérive des caractéristiques de transistors en fonction de la température.

Réaction et comportement de bruit par R. E. Bogner

Résumé de l'article aux pages 115 à 117

Le fait que le facteur de bruit et l'impedance de source optima d'un amplificateur sont indépendants de la réaction est prouvé par une méthode générale très simple. On en tire un moyen pratique de déterminer les faits sur la performance de bruits des circuits utilisés dans les voies de réaction. Des exemples de son application sont donnés.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 124 bis 129

Konstant-Stromversorgung

Kingshill Electronic Products Ltd, Torrens Works, Torrens Street, London, E.C.1

(Abbildung Seite 124)

Eine kompakte Stromversorgung mit Transistorstabilisierung für Labors mit einer in zwei umschaltbaren Bereichen von 0...50 V kontinuierlich regelbaren Ausgangsspannung bei maximal 1 A ist nunmehr von Kingshill Electronic Products Ltd erhältlich. Modell 501 hat für Quelle wie auch Verbraucher umfangreiche Schutzschaltungen. Diese vollelektronischen Schaltungen können entweder den Arbeitsstrom begrenzen oder den Ausgang bei acht einstellbaren Stromwerten zwischen 5 mA und 1 A völlig abschalten. Einstellung der Begrenzung und des Abschaltwertes erfolgt mittels Schiebeschalter, und Überlastung wird durch eine Signallampe angezeigt.

Bei 35° C Höchstumgebungstemperatur ist die Regelung 0,1 %, das Regelverhältnis 1000 und die Restwelligkeit unter 1 mV. Jeder Ausgangsschenkel hat zwei parallelgeschaltete Anschlussklemmen mit Schlitzen, um "wandernde Adern" zu verhüten.

Ein Messgerät mit Grossskalen für 0...25 V, 0...50 V und 0...1 A gestattet genaue Überwachung von Ausgangsspannung und -strom.

EE 78 751 *für weitere Einzelheiten*

Kernquadrupolresonanz-Spektrometer

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

(Abbildung Seite 124)

Das Scientifica - Kernquadrupolresonanz-Spektrometer wurde speziell entwickelt, um den Bedarf an solchen Ausrüstungen für Forschung und Vorführungszwecke bei wissenschaftlichen Abteilungen zu decken. Es ist für den Frequenzbereich 10...60 MHz ausgelegt und weist viele interessante Kernquadrupolresonanzen nach. Mit der mitgelieferten polykristallinen Standardprobe kann man bei einer Temperatur von 25,2° C mit Frequenzen von 34,233 und 27,007 MHz grosse Resonanzsignale von den beiden Chlor-Isotopen Cl^{35} und Cl^{37} erhalten.

Messen der Kernquadrupolresonanzfrequenzen in Proben, die mehr als eine Isotope enthalten, gibt einen direkten Vergleich der Kernquadrupolmomente der verschiedenen Isotopen und führt—für einige Moleküle—zur Bestimmung ihrer absoluten Werte. Für Kerne, deren Quadrupolmoment bereits bekannt ist, ermöglicht das Messen der Kernquadrupolresonanzen eine Voraussage der Elektronenverteilung innerhalb des Moleküls, und daraus ergibt sich Information über die chemische Grundstruktur der Probe. Man kann auf Grund dieser Information zum Beispiel entscheiden, ob eine bestimmte Bindung rein kovalent oder rein ionisch ist; bei einer gemischten Bindung lässt sich feststellen, zu welchem Prozentsatz jede der Typen beiträgt.

Obwohl in der reinen Kernquadrupolresonanz-Spektrometrie kein Magnetfeld erforderlich ist, kann man durch Anlegen eines schwachen magnetischen Feldes Zeeman-Spalten der Resonanzlinie hervorrufen. Messen des Spaltens in verschiedenen Probenrichtungen im Magnetfeld gestattet, die Abweichung des molekularen elektrischen Feldes von der Axialsymmetrie zu berechnen. Wenn die Probe einem inhomogenen Magnetfeld ausgesetzt wird, wird das Resonanzsignal durch sich ergebende Linienverbreitung verschmiert. Das kann sehr leicht vorgeführt werden, indem man den mitgelieferten kleinen Dauermagneten ungefähr 10 cm von der Probe entfernt positioniert.

Nachfolgend wird das Verfahren zum Erlangen von Kernquadrupolresonanzsignalen beschrieben: Die Probe wird in die Spule eines Detektoroszillators gelegt und der Oszillator kontinuierlich durchgestimmt. Wenn die Frequenz mit der für Resonanz erforderlichen übereinstimmt, absorbiert die Probe Energie vom Oszillator und ruft eine Änderung der Arbeitsbedingungen der Oszillatordetektorschaltung hervor. Es ist vorgesehen, dass die Frequenz laufend durch die Resonanzbedingung gewobbelt wird, woraus sich ein Absorptionssignal ergibt, das sich für Verstärkung in dem rauscharmen Transistorverstärker der Schaltung eignet. Das Ausgangssignal dieses Verstärkers kann direkt in einen Oszillografen mit niedriger Verstärkung gespeist und auf dessen Schirm dargestellt werden.

Eine Pendelfrequenz-Oszillatordetektorschaltung wurde entwickelt, um im Scientifica - Kernquadrupolresonanz - Spektrometer den grössten Signalabstand zu erlangen. Mittels eines Bedienelementes kann die Pendelfrequenz geändert werden, was gestattet, zwischen der Mittenresonanzfrequenz und ihren Seitenbändern zu unterscheiden. Ein weiteres Bedienelement ist mit "MOD. AMP." markiert und erlaubt Einstellung der Bandbreite, über die die Oszillatorfrequenz gewobbelt wird.

Ausser dem starren Messkopf für Aufnahme der Probe bei Zimmertemperatur wird auch noch ein etwa 30 cm langer beigsamer Messkopf mitgeliefert, mit Hilfe dessen die Probe entweder in einen Kühlschrank oder—für das Messen der Zeeman-Spaltung—in ein Magnetfeld gebracht werden kann. Zur Herstellung von Spulen nach Eigenwünschen für Proben unterschiedlicher Grösse oder für Grobnachstellung der Betriebsfrequenz wird eine Länge versilberten Drahtes mitgeliefert.

EE 78 752 *für weitere Einzelheiten*

Digitalkompensator

Scientific Furnishings Ltd, Poynton, Cheshire

(Abbildung Seite 124)

Der von Scientific Furnishings Ltd entwickelte neue Digitalkompensator kombiniert schnelle und bequeme Einknopfnullung mit den Vorteilen der Digitalanzeige der gemessenen Spannung. Die Kommastellung wird automatisch bei der Bereichumschaltung eingestellt, die Teilbereiche von 0...9,99 mV bis zu 0...499,9 mV überdeckt. Ein Einsteck-

Zusatz erweitert den Bereich bis zu 500 V mit Strombereichen von 0 ... 100 μA bis zu 5 mA. Mit einem Eisen-Konstantan-Thermoelement ist die Messempfindlichkeit besser als 0,2° C oder 10 μV mit 0,15 Prozent Absolutgenauigkeit.

Schnelles Nacheichen kann mittels des internen Weston-Normalelementes erfolgen, und eine 14-Ah-Quecksilberbatterie dient als Präzisionsspannungsquelle für allgemeine Eich- und Testzwecke bis zu 1 000 V. Für den Aussendienst, für den dieses kompakte Gerät (140 × 146 × 197 mm) sich besonders eignet, gibt es eine Ledertragtasche. Eine Halbleiterschaltung schützt das stossfeste Galvanometer gegen Überlastung.

EE 78 753 *für weitere Einzelheiten*

400-Hz-Stromversorgung

Hatfield Instruments Ltd, Burrington Way, Plymouth, Devon

(Abbildung Seite 125)

Hatfield Instruments Ltd hat eine neue dreiphasige transistorisierte 400-Hz-Stromversorgung herausgebracht, die in erster Linie für Hersteller von Präzisionskreiseln entwickelt wurde. Als 400-Hz-Oszillator wird in diesem Gerät ein Multivibrator mit ± 1% Konstanz benutzt, der mit einer externen Quelle synchronisiert werden kann. Ein von der Spannung im wesentlichen unabhängiger Phasenmesser ist vorhanden, was die genaue Einstellung von Verschiebungen bis zu 120° zwischen den drei Phasen sehr erleichtert. Eine 0,2-Prozent-Konstanz der Ausgangsspannung wird erzielt.

EE 78 754 *für weitere Einzelheiten*

Industrielle Steckverbindungen

Plessey-UK Ltd, Ilford, Essex

(Abbildung Seite 125)

Durch die allgemeine Entwicklung der Automation, Instrumentierung und Fernwirkung in der Industrie—besonders in der Schwerindustrie (z.B. Bergbau und Erdölindustrie)—stieg die Nachfrage nach zuverlässigen, robusten Steckverbindungen, die eine Anzahl von Kleinenergie- und Kleinsignalschaltungen aufnehmen können.

Die Steckverbindungen, Kupplungen, Kabel- und Rohrgarnituren der Baumuster 11 und 11F wurden von der Plessey-UK in enger Zusammenarbeit mit der britischen Nationalen Kohlenbehörde (NCB) entwickelt und entsprechen daher deren Anforderungen in bezug auf Höchstzuverlässigkeit und Widerstand gegen Umgebungseinflüsse, sowie Kleinstabmessungen und Kosten. Sie finden zur Zeit in den R.O.L.F.— und Collins-Kohlengewinnungsmaschinen sowie Förderanlagen und Fernsprechsystemen der Behörde Verwendung.

Die Konstruktion des Baumusters 11 beruht auf der der wohlerprobten Plessey-UK-AN-Steckverbindungen, von denen Hunderttausende unter schwierigsten Bedingungen in fehlerfreiem Betrieb arbeiten. In aus hochfesten Messing-Schmiedestücken oder -Stangen gefertigten Spezialgehäusen für Flantschenverbindung finden dieselben versilberten oder vergoldeten Kontakte für lötfreien oder gelöteten Leiteranschluss, dieselben aus Silikonkautschuk gepressten Isolierkörper Verwendung. Die Flantschenkonstruktion ermöglicht Einsatz der Stecker und Dose für Frontplattenmontage oder als freie Elemente mit oder ohne Kabeldichtungen und angebaute Rohrgarnituren.

Die vier Grundgrössen der Steckverbindungen nehmen die verschiedensten für 12, 24. 50 und 100A bemessenen Kontakttypen auf und können von 2- bis 48polig geliefert werden. Polarisationsstege und -nuten verhindern falsches Kuppeln, und Kreuzkuppeln ähnlicher Steckverbindungen wird durch Einsatz orientierter Ausführungen vermieden. Letztere haben in jedem Flantsch sechs Gewindelöcher. in die der Benutzer in einer beliebigen von 20 Kombinationen drei Stifte einschrauben kann, so dass die Stecker nur mit den komplementären Dosen zusammenpassen.

Dichtungen, die sich an Stecker und Dosen anschrauben lassen, sind speziell für die Aufnahme einer grossen Auswahl bewehrter oder unbewehrter Kabel in jeder Grundgrösse ausgelegt. Sie sind wasserfest, dichten und klemmen getrennt die Kabelisolierung und—wo erforderlich—auch die Bewehrung fest. Rohrgarnituren klemmen Schlauchleitungen und biegsame Leiter an die Steckverbindungen.

Die abgebildete Steckverbindung und Dichtung Baumuster 11F ist die flammenbeständige Ausführung des Baumusters 11. Sie entspricht den Anforderungen der britischen Norm BS 229 für flammenbeständige elektrische Apparate und wurde vom Bergbau-Institut Buxton unter Nummern FLP 5 344 und 5 345 für Einsatz mit Gasen der Gruppen I, II und III zugelassen.

EE 78 755 *für weitere Einzelheiten*

Akustische Anzeiger

Highland Electronics Ltd, 26-28 Underwood Street, London, N.1.

(Abbildung Seite 125)

Der akustische Sonalert-Festkörperanzeiger in Gleichstromausführung ist jetzt auch für 200 ... 250 V$\sim$ lieferbar.

Diese Bausteine geben ein auffallendes, leicht erkennbares akustisches Alarmsignal hoher Frequenz ab; sie sind klein, leicht, in sich geschlossen und für Einlochmontage ausgelegt. Sie enthalten keine beweglichen Kontakte, erzeugen keine elektrischen Störungen, sind höchstzuverlässig und für schwierige Umgebungsbedingungen konstruiert. Die Bausteine haben einen Durchmesser von 42,9 mm und sind ca. 57 mm lang.

Diese Sonalerts haben umfangreiche Anwendungsmöglichkeiten als Anzeiger, Signalgeräte, in der Überwachung, als Ruf- und Warneinrichtungen.

EE 78 756 *für weitere Einzelheiten*

Rückführungs-Drehmelder

Moore, Reed & Co. Ltd, Woodman Works, Durnsford Road, London, S.W.19

(Abbildung Seite 125)

Moore, Reed & Co. Ltd hat die Serienfertigung eines neuen kompensierten 400-Hz-Funktionsdrehmelders Grösse 11 Typ 11 RSF für Präzisionsrechneranlagen aufgenommen.

Für diese Drehmelder werden bei einem Fehler von unter 4 Minuten zwischen den Wellen Abweichungen von unter 0,05 Prozent von der Sinuswellenform angegeben, und in der Fertigung wurde ein bemerkenswert hoher Anteil von Geräten mit Abweichungen von unter 0,03 Prozent und Grundrauschpegeln von 1 ... 3 mV erreicht. Das Übersetzungsverhältnis zwischen Stator und Kompensationswicklung ist 0,97.

Trotz der hohen Genauigkeit und Allgemeinqualität sind die mit freien Zuleitungen oder 12-poligem Anschlussblock ausgerüsteten Einheiten billig lieferbar.

EE 78 757 *für weitere Einzelheiten*

Phasenschiebergenerator

Feedback Ltd, Crowborough, Sussex

(Abbildung Seite 126)

Der Phasenschiebergenerator VPO 230 ist ein Beispiel für die Anwendung der Funktionsverstärkertechnik auf einen Grossbereich-Sinuswellengeber.

Besondere Eigenschaften dieses Gerätes sind seine drei getrennt regelbaren Ausgänge: Bezugsphase, um 90° verschobene Phase und veränderliche Phase. Einmal eingestellt, werden sie über den Frequenzbereich 1Hz...100kHz mit konstanter Amplitude abgegeben. Die Ausgangsimpedanzen sind 600Ω.

Der Ausgang Bezugsphase: Quadraturphase wird innerhalb 0,25° von 90° gehalten, und die veränderliche Phase kann auf jeden gewünschten Winkel innerhalb 1° eingestellt werden.

Bei Betrieb als Oszillator ist der Klirrfaktor des Ausgangs niedriger als 0.5 Prozent.

Das für die Phasen- und Amplitudenmessungen angewandte Verfahren erfordert einen Oszillografen, an dessen horizontalen Ablenkkanal das phasenveränderliche Signal gelegt wird.

Der Vertikalkanal wird zum Messen der Amplitude verwendet, während der Phasenwinkel direkt auf der Phasenskala des Generators abgelesen werden kann, wenn die Lissajoussche Darstellung auf Null gebracht wird.

Ausserordentlich erhöht wird die Vielseitigkeit durch eine externe Mitnahmeeinrichtung, mit der das Instrument innerhalb seines Bereiches als "Tochtergerät" jeder anderen Frequenzquelle arbeiten kann, was Phasenmessungen an Systemen ermöglicht, die mit höherer Leistung betrieben werden.

Bei Betrieb als "Tochtergerät" verhält sich das Instrument wie ein frequenzempfindlicher Verstärker und lässt sich so abstimmen, dass es harmonische Komponenten eines komplexen Hauptsignals selektiert.

EE 78 758 *für weitere Einzelheiten*

Temperatur-Prüfschrank

Montyford Instruments Ltd, 79 Lots Road, London, S.W.10

(Abbildung Seite 126)

Der Prüfschrank hat einen Bereich von $-70°$ C ... $+200°$ C und erlaubt jähe Temperaturänderung, die man von der Frontplatte aus regeln kann. Ausser der Temperaturänderung mit regelbarer Geschwindigkeit kann durch einen Umschalter auf periodisches Schwanken zwischen zwei vorgewählten Temperaturen oder hohe Konstanz bei einer Temperatur umgestellt werden.

Programme lassen sich zu jedem gewünschten Zeitpunkt stoppen oder umkehren. Kühlung erfolgt mittels CO_2, dessen Einblasen in den Wärmeaustauscher durch einen elektrischen Zeitgeber gesteuert wird. Die Heizung ist in den zwangsläufigen Luftumlauf eingebaut, was örtliche Überhitzung vermeidet und Temperaturgefälle gering hält.

Kammer und Bedienfeld sind in Modultechnik ausgeführt und in einem Standardgestell untergebracht, das sich für den Einbau von Programmsteuerungen, Verstärkern und anderen Ausrüstungen für Gestelleinbau, die für die zu prüfenden Bauelemente oder Geräte gebraucht werden, ausbauen lässt.

Da elektrische Verbindungen in die Prüfkammer hineingebracht werden können, wird dynamisches Testen von Bauelementen und Geräten ermöglicht.

EE 78 759 *für weitere Einzelheiten*

Drehmomentgeber

Mechatronics (London) Ltd, 317 Kennington Road, London, S.E.11

Mechatronics (London) hat einen neuen Drehmoment-Messwertwandler eingeführt.

Dieser Messwertwandler wurde speziell für das Prüfen von Kleinstmotoren, hydraulischen Pumpen und Motoren, Generatoren und Alternatoren entwickelt und ist im Grunde genommen eine Drehmoment-Messapparatur, die das für ein Reaktionsdynamometer herkömmliche Lager überflüssig macht.

Die Apparatur bildet eine Auflage für die Krafteinrichtung sowie die ein von dieser Einrichtung entwickeltes Reaktionsdrehmoment in ein proportionales elektrisches Signal umwandelnde Anlage und besteht aus einer Platte, die mittels drehmomentempfindlichen Federn über einer feststehenden Grundplatte angeordnet ist. Auf diese Federn geklebte elektrische Dehnstreifen bilden eine Dehnstreifenbrücke.

Die Federn sind in Winkeln von 45° aufmontiert. Die Drehachse der Plattform wird durch den Schnittpunkt einer durch die Biegefedern gezogenen Linie dargestellt. Jede Apparatur, die so auf die Plattform gespannt ist, dass ihre Eingangs- oder Ausgangswelle genau mit der Drehachse zusammenfällt, wird ein Reaktionsdrehmoment hervorrufen, das von der Plattform wahrgenommen wird.

In Abwesenheit von Lagern und Schleifringen hat die Drehmomentplattform weder Reibungsverluste noch Drehzahlgrenzen. Man kann sie daher ohne Reibungsfehler für sehr niedrige Drehmomente verwenden.

Der Brückenwiderstand ist 100 Ω und die Brückenspannung 12 V; der Nennausgang ist 2 mV/V. Für Bereiche von 0...1132 g.cm bis zu 0...113 kg.cm lieferbar.

EE 78 760 *für weitere Einzelheiten*

Digital-Servokomparator

Epsylon Industries Ltd, Faggs Road, Feltham, Middlesex

(Abbildung Seite 126)

Die Funktion dieses Komparators ist das Vergleichen von zwei Frequenzen und die Erzeugung eines ihrer Differenz proportionalen Ausgangs in Form von Impulsen, deren Frequenz der Differenz zwischen den Originalfrequenzen entspricht, oder von einer Analogspannung, die der Frequenzdifferenz proportional ist.

Grundsätzlich werden die Eingangssignale erst nach Amplitude und Zeit in Rechteckwellen umgewandelt und dann jeweils ein kompletter Impuls der höheren Frequenz für einen Impuls der niedrigeren Frequenz abgezogen, und zwar unabhängig von der Phasenbeziehung zwischen den beiden Signalen.

Die Ausgangswellenform ist eine Impulsfolge, in der die Impulslänge $\frac{1}{2}f_1$ und die Impulsfrequenz $f_1 - f_2$ ist, die einen Digitalausgang ergeben. Die gemittelten und geglätteten Impulse geben einen Analogausgang.

Der Vorteil dieser Anordnung besteht darin, dass keine Zeitschaltung benötigt wird, die Genauigkeit also nur noch von der Ansteigszeit der Ausgangsschaltung abhängt. Das Gerät kann durch Zusatz weniger Elemente, die die Schaltung differential gestalten, noch vielseitiger gemacht werden. In der ersten Anordnung wird der Ausgang für $f_2 > f_1$ Null. In der letzteren Ausführung wird der Ausgang als $|f_1 - f_2|$ und mit einer Kennzeichnung des Signals mit der höheren Frequenz gegeben. Die Digitaldarstellung erfordert nur Ausblenden des Ausgangs in festen Zeitabständen in ein Zählwerk. Andererseits kann das Analogausgangssignal direkt auf einem Voltmeter angezeigt werden.

Für Einsatz zur automatischen Geschwindigkeitsregelung kann Leistungsverstärkung bis zu 2,5 kW vorgesehen werden.

Die Grundfrequenz des Gerätes ist 0 ... 10 MHz, und die Eingänge können zwischen 3 und 100 V liegen; die Eingangsimpedanz ist 10 kΩ. Der Ausgang ist 6 V bei 1 mA (digital) oder 0 ... 6 V bei 1 mA (analog).

EE 78 761 *für weitere Einzelheiten*

Halbleiterprüfer

Vacwell Engineering Co. Ltd, Sherman Road, Bromley North, Kent

(Abbildung Seite 126)

Vacwell Engineering Co. Ltd kündigt ein neues Gerät in ihrem Fertigungsprogramm für die Halbleiterindustrie an. Es handelt sich um den "Laboratoriums-Scheibenprüfer" PR.61, der in erster Linie für Stichproben von Transistorplättchen, die noch in Scheibenform sind, bestimmt ist. Doch hat er auch andere Anwendungsmöglichkeiten, vor allem in der Mikroschaltungstechnik.

Das für Tischmontage ausgelegte Gerät wird üblicherweise mit zwei Prüfköpfen geliefert, jedoch kann die Anlage bis auf vier Prüfköpfe erweitert werden. Diese Köpfe sind elektrisch vom Hauptkörper isoliert und werden in Betrieb mittels eines kleinen Handhebels angehoben und auf die Scheibe gesenkt. Fehlerhafte Scheiben kann man in der Prüfposition mittels eines drucktastengesteuerten Auftragkopfes farbkennzeichnen.

Die zu prüfende Scheibe wird in einem isolierten Vakuumfutter, das zur Ausrichtung der Scheibe drehbar ist, auf eine X-Y-Plattform mit Mikrometereinstellung eingespannt.

Das Pflichtenblatt des PR.61, der ohne Schwierigkeiten mit einem halbautomatischen Vacwell-Transistortester für drei Parameter oder andere Prüfausrüstungen gekuppelt werden kann, umfasst auch ein Binokularmikroskop mit 50facher Verstärkung und Beleuchtung der Arbeitsplattform. Während des Einspannens der Scheibe lässt sich der Mikroskopzusatz ausschwenken.

EE 78 762 *für weitere Einzelheiten*

Subminiatur-Zungenschalter

Flight Refuelling Ltd, Wimborne, Dorset

(Abbildung Seite 127)

Flight Refuelling Ltd kann nunmehr Hamlin-Zungenschalter "A" in der neusten Ausführung liefern. Die Konstruktion des neuen Subminiatur-Zungenschalters MTRG-2 entspricht der des vor kurzem eingeführten MRG-DT des Baumusters "C". Weder der MTR-DT, noch der MTRG-2 erfordert magnetische Vorspannung, und da der Kontaktabstand in beiden Schaltern fluchtet, gibt es in dieser Kleinstgrösse zweipolige und einpolige Umschalter sowie einpolige Kippschalter.

Bei einem Glasdurchmesser von 3,55 mm ist die Gesamtlänge 54 mm; die Zuleitungen können zum Packen auf 19 mm geschnitten werden. Diese Schalter haben von der Standardausführung alle klimatischen Eigenschaften übernommen; bei Ohmscher Vollast von 3 W—(28 V_{max}, 0,250 A_{max}) haben sie eine erwartete Lebensdauer von 10 × 10^6 Vorgängen. Der Kontaktwiderstand hat einen Anfangshöchstwert von 0,150 Ω und einen Nennwert von 0,100 Ω.

Zum Vergleich der Abmessungen zeigt die Abbildung den Subminiatur-Zungenschalter MTRG-2 neben der Standardgrösse DRG-2 Baumuster 'A'.

EE 78 763 *für weitere Einzelheiten*

Temperaturregler

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(*Abbildung Seite 127*)

Die Fertigung einer Serie von Festkörper-Temperaturreglern modernster Konstruktion, die Temperaturen bis zu 2000° C in elektrischen Öfen mit 0,5 Prozent Genauigkeit regeln, ist bei Kent Precision Electronics Ltd angelaufen. Das System besteht aus einem herkömmlichen Temperaturwandlerelement, das ein der Temperatur proportionales Signal in einen genauen Transistor-Gleichstromverstärker speist. Das Eingangssignal wird mit einem konstanten und genauen Bezugssignal verglichen, dessen Grösse auf der Anzeigeskala auf der Frontplatte nach Wunsch eingestellt wird; das Fehlersignal steuert das Thyristor-Ausgangssystem, das die an die Heizelemente abgegebene Leistung regelt.

Für niedrige Leistung wird die proportionale Regelung der Ausgangsleistung durch Änderung des Phasenwinkels, bei dem die Thyristorschaltung während jeder Halbperiode der Netzwechselspannung zündet, erzielt. Für hohe Leistungen wird die Thyristorschaltung für ganze Halbperioden eingeschaltet und die proportionale Regelung durch Änderung des Verhältnisses der Anzahl ein- und abgeschalteter Perioden erreicht. In allen Fällen ist die proportionale Regelung über ungefähr 3 Prozent der Skala wirksam.

Für die Regelung von Strömen über 100 A—wenn die Thyristorregelung unwirtschaftlich wird—können Standard-Thyristorausgangsschaltungen für niedrige Leistungen eingesetzt werden, die magnetische Verstärker für proportionale Regelung oder elektromagnetische Kontaktgeber für Ein-Ausbetrieb steuern. Das Foto zeigt eine Standard-10-kW-Temperaturregleranlage für proportionale Regelung.

EE 78 764 *für weitere Einzelheiten*

Miniatur-Einstellpotentiometer

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

Typ 62 ist ein neues, von Morganite Resistors Ltd hergestelltes Einstell-Potentiometer mit Schraubenzieherregelung, das speziell für Miniaturgeräte, in denen kleine Abmessungen von ausschlaggebender Bedeutung sind, bestimmt ist. Die Abmessungen des Bauelementes sind etwa 12 × 9 mm.

Der Lötfahnenabstand entspricht einem Rastergrundmass von 2,5 mm. Der Schleifer ist von beiden Seiten des Bauelementes aus einstellbar, und die Widerstandsbahn ist während des Einstellens gegen zufällige Beschädigung geschützt.

Zwei Ausführungen werden angeboten, beide für Bestückung gedruckter Schaltungen. Typ 62H ist für waagerechte Montage ausgelegt.

Die Nennleistung ist bei 40° C Umgebungstemperatur 0,2 W und der Widerstandsbereich 100 Ω ... 1 MΩ.

EE 78 765 *für weitere Einzelheiten*

Drehmomentgeber

Saunders-Roe Division, Westland Aircraft Ltd, Yeovil, Somerset

(*Abbildung Seite 127*)

Sowohl Rotor- wie auch Statorsatz des neuen Drehmomentgebers Typ 2 Baumuster 4 sind in einem Stück in Glasfaserkonstruktion gepresst. Das Drehmoment wird durch eine hochfeste Stahlwelle übertragen, auf der Saunders-Roe-Filmdehnstreifen aufgebracht sind. Elektrische Verbindungen mit Rotor und Stator werden durch Silberschleifringe und Silber-Kohlebürsten erstellt und dann über ein vieradriges angepresstes Kabel der Instrumentierungsausrüstung zugeführt. Der gesamte, auf Linearität, Hysterese und Wiederholbarkeit zurückführbare Fehler liegt innerhalb ±0,2%.

Für die Instrumentierung dieses Messwertwandlers steht der Anzeiger TM6 zur Verfügung, der für jeden beliebigen Wandler zwei volle Drehmoment-Skalenbereiche hat. Ausser der direkten Anzeige des echten Mitteldrehmomentes auf der Kreisskala sind Buchsen für den Anschluss zusätzlicher Schreiber, Oszillografen und anderer wahlweiser Zusätze vorhanden. Die Anzeigegenauigkeit des echten Mitteldrehmomentes wird durch grosse periodische Schwankungen des Drehmomentes nicht beeinflusst.

Der Geschwindigkeitsbereich des Systems überstreicht 0 bis zu der der Grösse der Messwandlers entsprechenden maximalen Nenngeschwindigkeit; der Frequenzgang ist sehr günstig für Einschwing- und periodische Untersuchungen. Der Anzeiger kann entweder ans Wechselstromnetz angeschlossen oder aus Batterien gespeist werden. Für Bereiche von 0 ... 0,7 mkg bis zu 0 ... 6900 mkg und Geschwindigkeiten von 0 ... 8000 UPM bis zu 0 ... 800 UPM sind 17 Messwandlergrössen lieferbar.

EE 78 766 *für weitere Einzelheiten*

Vorwählzähler

Elesta Electronics Ltd, 34 Cambridge Park, London, E.11

(*Abbildung Seite 128*)

Dieses kompakte, tragbare Gerät kann bis zu 100 kHz zählen. Es ist in Einsteck-Druckkartentechnik ausgeführt und durchweg mit Silizium-Elementen bestückt.

Ein Knopfschalter gestattet Vorwahl einer Zählfolge oder Abzählen von Gegenständen in Partien; das Gerät ist mit vier Langlebensdauer-Zählröhren bestückt, die eine klare und helle Anzeige geben. Der Zähler selbst kann ohne Zähldekadenanzeige, aber mit Vorwahl, gegen einen entsprechenden Preisabschlag geliefert werden.

Die folgenden Konstruktionsmerkmale des Gerätes gestatten Betrieb bei unterschiedlichen Arbeitsweisen:

(1) Automatische Rückstellung auf Null, sowie die vorgewählte Zeit erreicht ist. Man kann das Ausgangsrelais für eine zwischen 20 und 200 ms einregelbare Zeitspanne geschlossen halten.

(2) Bei einfachem Abzählen kann manuell zurückgestellt werden.

(3) Bei Einstellen auf automatische Rückstellung ist die Anzeigezeit zwischen 20 ms und 2 s regelbar.

Modulare Einsteckschaltungen ermöglichen es, komplette Systeme für Zählen, Abzählen von Partien oder Regelung nach Kundenwünschen zu bauen. Das anpassungsfähige System erlaubt auch Einbau einer fünften Zähldekade mit Vorwahlschalter und Höchstzählung von 99999 ohne Vergrösserung des Gehäuses.

Ein interner Schmitt-Trigger gewährleistet, dass der Zähler mit einer breiten Auswahl von Eingangswellenformen arbeiten kann. Frequenznormale für 4, 5, 6, 8 oder 10 Hz können als Sonderzubehör geliefert werden.

EE 78 767 *für weitere Einzelheiten*

Ultraschall-Regelgenerator

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(*Abbildung Seite 128*)

Der Soniclean-Automatic 1191B, ein Ultraschall-Regelgenerator, kann jetzt von Dawe Instruments Ltd bezogen werden. Er wurde für Verwendung mit einer Anzahl von Reinigungsbehältern wie z.B. dem Dawe Typ 1165/126/8, ein 22,7-l-Modell mit stark abgerundeten Ecken mit acht Schwingungswandlern und einer wirksamen Reinigungsfläche von etwa 806 cm^2, entwickelt und besteht aus einem Transistor-Oszillator mit transformatorgekoppelter Gegentaktendstufe. Die Endstufe ist ihrerseits über einen Transformator mit der Schwingungswandlerlast gekoppelt, die ein integraler Teil des Standard-Behälters ist. Eine Gegenkopplungsschaltung regelt die Eingangsspannung des Verstärkers gemäss dem Strom in den Bewegungsarmen des Schwingungswandlers und bewirkt, dass der Generator dauernd die Resonanzfrequenz des Wandlers sucht und hält, und zwar unabhängig von Behälterbelastung, Temperatur oder anderen Betriebsbedingungen. Die einzigen erforderlichen Bedienelemente sind daher ein einfacher Ein-Ausschalter und eine Warnlampe.

Die meisten derzeitigen Ultraschall-Reinigungssysteme werden manuell auf den optimalen Arbeitspunkt, d.h. die Resonanzfrequenz des Schwingungswandlers, abgestimmt. Diese Anordnung arbeitet bei der Reinigung im kleinen Massstab völlig zufriedenstellend, ist aber für viele Betriebsbedingungen zu langsam und verlangt eine gewisse Erfahrung. Das Regelsystem von Typ 1191B gewährleistet andererseits automatisch jederzeit wirtschaftliches Arbeiten.

Da der Wandler durch die laufende Einregelung auf Resonanz Beanspruchungen unterliegt, findet ein neuer Schwingungswandlertyp Verwendung, dessen aktive Elemente aus Bleizirkonattitanat statt aus dem üblicheren Bariumtitanat bestehen. Die neuen Wandler Typ 1166 werden mit den Standardbehältern geliefert, sind jedoch auch getrennt erhältlich. Sie haben einen besonders hohen Wirkungsgrad und können bei Temperaturen bis zu 149° C betrieben werden, was für alle voraussehbaren Reinigungsvorgänge ausreicht. Beheizte Ausführungen des Standardbehälters mit Temperaturregelung können ebenfalls geliefert werden.

Ultraschallenergie wird im Typ 1191B in Impulsen bei der Resonanzfrequenz des Systems (ca. 25 kHz) und mit einer Folgefrequenz von 100 Hz oder der doppelten Netzfrequenz, die durch Vollweggleichrichtung erreicht wird, erzeugt. Die Spitzen-Ausgangsleistung ist 600 W, der Mittelwert etwa 300 W, und dem Netz werden 600 W entnommen. Der Generator ist volltransistorisiert und wird für Tischmontage in einem Metallgehäuse 406 mm × 356 mm × 165 mm hoch geliefert; er wiegt etwa 18 kg.

EE 78 768 *für weitere Einzelheiten*

Abstimm-Varactor hoher Güte

Microwave Associates Ltd, Cradock Road, Luton, Bedfordshire

Microwave Associates Ltd hat vor kurzem zwei neue Serien von Silizium-Abstimmvaractoren mit hohem Q vorgestellt.

Eine Serie legierter Epitaxial-Abstimmvaractoren bietet einen Mindest-Q von 100 bei −4 V, 50 MHz, mit einem charakteristischen Kapazitätsverhältnis von 5,2 : 1 für eine Vorspannungsänderung von 2 ... 100 V−. Lieferbar in Kapazitätswerten von 6,5 bis 47 pF bei −4 V und einer Spitzensperrspannung von mindestens 110 V.

Q-Werte bis zu 350 bei −4 V, 50 MHz stehen in einem breiten Sortiment diffundierter Flächenvaractoren für Abstimmzwecke zur Verfügung. Die Spitzensperrspannung ist 35 V mit einem Kapazitätsverhältnis von 3 : 1 zwischen 2 und 30 V. Dioden für höhere Spannungen können auf Wunsch geliefert werden.

Beide Serien kommen in einem Subminiatur-Glasgefäss mit Anschlussdrähten, und die Mikropille (1,4 mm hoch) wird auch für Kapazitätswerte von 15 pF und niedriger angeboten. Die Höchstbetriebstemperatur ist 150°C mit einer Verlustleistung von 600 mW.

EE 78 769 *für weitere Einzelheiten*

Steckbares Relais

Simmonds Relays Ltd, Edinburgh Place, Harlow, Essex

(*Abbildung Seite 128*)

Simmonds Relays Ltd kündigt zwei Steckausführungen ihres Relais Typ 3 000 an. Sie entsprechen den Anforderungen der meisten Verwendungszwecke, in denen völlig gekapselte, staubdichte Relais dieser Type benötigt werden.

Beide Typen sind von einer durchsichtigen Polykarbonathaube umschlossen, und in beiden Fällen werden die Relais durch Halteklammern aus rostfreiem Stahl in ihren Fassungen festgehalten.

In der Ausführung 3 000/SP werden die Fahnenenden speziell ausgelegter kurzer Kontaktfedern als Verbindungsmittel benutzt. Eine Phenolharz-Pressfassung, in die das Relais einsteckbar ist, wird mitgeliefert. Dieses Relais ist besonders für Einbau in die Standardgestelle der britischen Postbehörde konstruiert.

In der anderen Ausführung, Typ 3 000/20P, sind Kontaktfedern und Spulenenden intern direkt an eine rechteckige Sockelpressplatte mit 20 Stiften angeschlossen, die fest mit dem Relais verbunden ist. Auch wird mit jedem Relais eine entsprechende Fassung mitgeliefert.

Beide Relais können für bis zu vier hochbeanspruchte oder bis zu acht leicht beanspruchte Aktionen bestückt werden. Die Kontaktdaten sind genau wie für das Standardrelais 3 000.

In dem einen oder anderen der obigen Gehäuse untergebrachte Sonderausführungen sind lieferbar, u.a. Relais mit Mikroschaltern und offenen Kontakten, Relais mit Kontaktisolierung bis zu 10^{13} MΩ, Kupfermantelrelais, Haftrelais, Relais mit nickeleisenummantelten Kernen und hoher Impedanz bei Wechselstrom, Relais mit bifilar-, dreifach- oder scheibengewickelten Spulen.

EE 78 770 *für weitere Einzelheiten*

Elektrische Stoppuhr

Venner Ltd, Kingston By-Pass, New Malden, Surrey

(*Abbildung Seite 129*)

Venner Ltd hat eine elektrische Stoppuhr eingeführt, die ein neues und genaues Hilfsmittel für Physikklassen bis zur Grundschule herunter darstellt. Es ist nicht mehr nötig, Daten auf komplizierte Weise zu verarbeiten, wie es z.B. bei allen früheren Verfahren zur Bestimmung der Beschleunigung durch Schwerkraft oder zur Untersuchung der beschleunigten Bewegung einer Laufkatze bedingt war.

Die ursprünglich vom Oberlehrer für Physik am Trinity College in Glenalmond entwickelte Stoppuhr hat eine einfache Arbeitsweise mit elektrischem Start/Stopp und elektrischer oder manueller Rückstellung. Sie wird von einem für 125 V~, 50 Hz Netzanschluss bemessenen Synchronmotor getrieben. Kontakte geben jede Sekunde einen Ausgangsimpuls, und ausserdem können kontinuierlich 12 V− entnommen werden.

Die Hauptskala hat einen Durchmesser von ca. 63,5 mm mit weisser Beschriftung auf schwarz, ist in 1/100tel Sekunden unterteilt und wird jede Sekunde einmal von einem grossen Zeiger durchlaufen. Von zwei kleinen Zeigern durchläuft der eine eine innere Skala einmal in 10 Sekunden, der andere einmal in 150 Sekunden; die 10-s-Skala hat 1-Sekunden-, die 150-s-Skala 10-Sekunden-Unterteilungen.

Die Stoppuhr ist so genau wie die Netzfrequenz.

Um das Aufeinanderstapeln von Uhren zu erleichtern, sind Anschlussbuchsen und Tasten für manuelles Rückstellen unter der oberen Kante des berührungssicheren Gehäuses angeordnet.

EE 78 771 *für weitere Einzelheiten*

Koaxiale Adapter

Greenpar Engineering Ltd, Station Works, Cambridge Road, Harlow, Essex

(*Abbildung Seite 129*)

Adapter für die gegenseitige Anpassung von Steckern und Buchsen der Serien "BNC", "C", "N" und "UHF" in jeder beliebigen Kombination sind aus einem neuen Fertigungsprogramm lieferbar. Sie können entweder als Bausätze mit zwei Schlüsseln und Kästchen für Verwendung im Labor oder Aussendienst, oder als dichte Ausführung für Standardverwendungszwecke geliefert werden. Der Bausatz ermöglicht die Erstellung von vier beliebigen kompletten Adaptern in 28 möglichen Kombinationen.

Jedem Element dieses Programmes wurde eine NATO-Katalognummer zugeteilt, und sie sind gegen jeden nach US-Militärzeichnungen gefertigten äquivalenten Adapter zur gegenseitigen Anpassung zwischen Serien austauschbar.

Mit Ausnahme der Stecker und Buchsen der Serie "UHF" (die nicht angepasst sind) ist das Stehwellenverhältnis wesentlich verbessert; bei 4 GHz ist es besser als 1,1.

Alle internen Kontakte sind vergoldet.

EE 78 772 *für weitere Einzelheiten*

Sechskanal-Verstärker

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(Abbildung Seite 129)

Das neue Sechskanal-Verstärkersystem R6 ist für die Mehrkanalverstärkung von schwachen Signalen bestimmt, wie sie z.B. von Thermoelementen, Dehnstreifen und Messwertwandlern abgegeben werden. Die sechs Verstärker sind in einen 19"-Einschub eingebaut. Das Netz wird von hinten angeschlossen, Eingang und Ausgang auf der Frontplatte. Die Verstärkung kann für jeden Verstärker einzeln zwischen 330 und 2 400 geregelt werden. Die Fotowiderstand-Chopperverstärker haben mit unter 0,2 μV/°C eine sehr niedrige Drift. Der hohe Eingangswiderstand von 100 MΩ verhindert Belastung der Quelle; mit ihrem hohen Ausgangsstrom von 25 mA und einer Bandbreite von 20 kHz eignen sich die Verstärker für das Treiben ultravioletter Schreiber.

EE 78 773 *für weitere Einzelheiten*

UKW-UHF-Empfänger

VHF/UHF Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Abbildung Seite 129)

Der UKW-UFH-Empfänger TR 40/20 ist ein tragbares und volltransistorisiertes Gerät für den Empfang von AM-Sprachübertragungen.

Er ist für Einkanalbetrieb im UHF-Band 430 ... 470 MHz gebaut und kann ausserdem in einem Einzelkanal des UKW-Bandes 70 ... 90 MHz betrieben werden. Er ist für den niedrigen Kanalabstand von 25 kHz gedacht und hat für UHF eine dreifache, für UKW eine zweifache Überlagerungsschaltung. Drei Stufen sind quarzgesteuert.

Ein kleiner Lautsprecher ist vorhanden, und eine interne 8-V-Quecksilberbatterie erlaubt etwa 70 Stunden Betrieb. Die äusserst niedrige Stromaufnahme ist auf die 18 Transistoren und vier Dioden zurückzuführen.

Man kann entweder eine steckbare Peitschenantenne oder ein externes 75-Ω-Antennensystem benutzen. Der TR 40/20 wiegt etwa 1,8 kg, ist 203 mm hoch, 178 mm breit und 51 mm tief.

Ein volltransistorisierter Sender niedriger Leistung Typ TT 40 wird zur Zeit für Einsatz mit diesem Empfänger entwickelt, ebenso ein kombiniertes Sende-Empfangsgerät TRT/40 für Simplex- oder Duplexbetrieb.

EE 78 774 *für weitere Einzelheiten*

Druckgeber

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

Der Geschäftsbereich Consolidated Electrodynamics der Bell & Howell Ltd hat einen neuen Druckgeber 4-390 mit grossem Ausgangssignal eingeführt, der besonders für zuverlässiges Arbeiten in extremen Beschleunigungs-, Schwingungs- und Stossbedingungen konstruiert wurde.

Das Modell 4-390 besteht aus einem nicht aufgekitteten Dehnstreifen als Messelement, mit integralem Festkörperverstärker und Stromversorgung; es gibt ein Ausgangssignal von 5 V— ab. Der Wandler ist für absoluten Druck und Überdruck von Flüssigkeiten und Gasen, die mit rostfreiem Stahl 416 vereinbar sind, konstruiert und für Druckbereiche von 7 kg/cm² bis zu 350 kg/cm² lieferbar.

Der Wandler hat weder Schleifkontakte, noch andere mechanische Teile, die sich bei Schwingungen abnutzen; die Stromversorgung gestattet beträchtliche Schwankungen der Eingangsleistung ohne Verschlechterung der Datensignale.

Als weitere Konstruktionsmerkmale sind ein Tiefpass zur Unterdrückung des hochfrequenten Rauschens im Ausgang, sehr gute Isolation zwischen Energieeingang und Signalausgang sowie eine Einrichtung, die Überschreiten eines vorgewählten Pegels durch das Ausgangssignal verhindert, zu nennen.

Das Instrument ist voll gegen zufällige Polaritätsumkehr der Eingangsenergie geschützt, und elektrische Bauelemente sind als integrale Einheit temperaturkompensiert, wodurch Temperatureinflüsse klein gehalten werden.

Der Messwertgeber ist für elektrische Erregung mit 28 V— bei Höchststromaufnahme von 50 mA ausgelegt. Linearität und Hysterese sind ±0,5 Prozent des Bereichendwertes, Ausgangsimpedanz niedriger als 250 Ω und das Gewicht etwa 285 g.

EE 78 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Eine Fading-Maschine zur Nachbildung der Ionosphäre von D. Beck und J. A. Betts

Zusammenfassung des Beitrages auf Seite 74-79

Die Fading-Maschine bildet die regellosen Schwankungen der Amplitude und des Phasenwinkels eines HF-Weitverkehrfunksignals durch Addition der Ausgänge von acht Quarzoszillatoren bei nominell 100 kHz nach. Gleichzeitige Ausbreitung über zwei Wege mit bis zu 2 Millisekunden Wegzeitdifferenz ist berücksichtigt. In der derzeitigen Ausführung gestattet die Maschine nur Einantennenbetrieb; Verdopplung der Ausrüstung würde dagegen Zweifach-Diversityempfang ermöglichen. Gaussches Weissrauschen kann dem Signal zugesetzt werden, um Signalabstände von 0 . . . 50 dB zu geben. Der Eingang kann entweder aus einem Funktelegrafiesignal oder einem beliebigen Einseitenbandsignal im Bereich 300 . . . 6000 Hz bestehen.

Der Feldeffekt-Transistor als Chopper bei niedrigem Pegel von K. Barton

Zusammenfassung des Beitrages auf Seite 80-83

Der Feldeffekt-Transistor-Chopper wird mit dem herkömmlichen Silizium-Legierungstransistor-Chopper verglichen. Es wird gezeigt, dass bei Verwendung eines Feldeffekt-Transistor (F.E.T.)-Choppers ohne aufwendige Selektionsverfahren oder Thermostaten eine Spannungsdrift von weniger als 0,1 μV/°C und eine Langzeitdrift von weniger als 0,1 μV pro Monat ohne weiteres erreichbar sind.

Die Anhänge bringen u.a. die Berechnungen für die Durchlassverstärkung des FET-Choppers, die EMK und Stromeingänge, sowie die Berechnung der auf Restströme zurückzuführenden Gegenspannungen.

Festwiderstände für hohe Spannungen von P. L. Kirby

Zusammenfassung des Beitrages auf Seite 84-89

Mittel, die durch Kontrolle der Filmwiderstandgeometrie den Spannungsgradient auf ein Kleinstmass reduzieren können, werden besprochen. Das Problem der Nichtlinearität und die Grenzen eines

"Spannungskoeffizienten" werden diskutiert. Für Verwendung bei hohen Spannungen ist es zweckdienlich, die Auswirkungen von Spannungs- und Temperaturkoeffizienten zu kombinieren. Ein Hochspannungswiderstand verbesserter Ausführung, der aus einem dicken Film aus "gestrecktem" Zinnoxyd besteht, wird beschrieben. Der Rauschpegel ist bemerkenswert niedrig; unter normalen Bedingungen haben die Widerstände eine Konstanz von etwa ± 1 Prozent und können Spannungsimpulse bis zu 50 kV_s vertragen. Als Schutz gegen hohe Spannungsstösse können Funkenstrecken vorgesehen werden. Das Verhalten dieser Widerstände in der Widerstandskette eines elektrostatischen Hochspannungsgenerators wird beschrieben.

Der Antennengewinn von Yagi-Richtstrahlfeldern von P. Knight und R. D. C. Thoday

Zusammenfassung des Beitrages auf Seite 91-93

Yagi-Querstrahlfelder finden häufig zur Verbesserung des Antennengewinnes und der Richtwirkung in UKW- und UHF-Bändern Verwendung. Wegen der komplizierten Form der Strahlungscharakteristik dieser Antennen ist jedoch das herkömmliche Verfahren zur Berechnung des Gewinnes, das auf der Integration der Strahlungscharakteristik der Leistung beruht, nicht ohne Schwierigkeiten anwendbar. Der Beitrag beschreibt ein alternatives Verfahren zur Berechnung des Antennengewinnes, das auf der Bestimmung des Gewinnwiderstandes zwischen Yagi-Paaren beruht. Weiterhin wird eine Tabelle gegeben, mit deren Hilfe sich der Antennengewinn einer Anzahl von Yagi-Richtfeldstrahlern leicht bestimmen lässt.

Ein Überblick über die Nachbeschleunigungsmethoden in Oszillografenröhren von K. J. Ming

Zusammenfassung des Beitrages auf Seite 94-97

Die Arbeitsweise der in gegenwärtigen elektronischen Geräten gängigsten Oszillografenröhren wird umrissen. Die meisten dieser Röhren haben als Nachbeschleunigungssysteme entweder einfache Bänder oder eine Graphitspirale.

Durch verbesserte, mit Gitter- oder Maschenelektroden ausgerüstete Nachbeschleunigungssysteme werden jetzt Röhren mit grösserer Ablenkempfindlichkeit lieferbar. Die charakteristischen Eigenschaften dreier solcher Systeme—der axiale Beschleuniger, der radiale Beschleuniger und der Hohllinsenbeschleuniger—werden beschrieben und verglichen.

Ein grafischer Analog-Digitalumsetzer von R. L. G. Gilbert und N. S. Oakey

Zusammenfassung des Beitrages auf Seite 98-101

Die in diesem Beitrag beschriebene Ausrüstung kann automatisch mit einem heissen Schreibstift erstellte Registrierstreifen lesen und die Information in Digitalform darstellen. Die Geometrie des Abtastsystems ist der eines Lichtpunktabtasters ähnlich, und die Position des Schriebs wird mittels eines Zähler-Zeitgebers gemessen. Der Ausgang kann mit jedem Digitalregistriergerät, z.B. einem Zeilendrucker, registriert werden. Die Unsicherheit ist $\pm 0{,}5$ Prozent des Endwertes und die Wiederholbarkeit $\pm 0{,}1$ Prozent.

Spurwahlsystem eines Magnettrommelspeichers von S. C. Acharyya

Zusammenfassung des Beitrages auf Seite 102-105

Der Beitrag befasst sich mit dem Entwurf eines Spurwahlschalters für ein Magnettrommel-Speichersystem. Durch Verwendung von Transistor-Torschaltungen im Schalter werden der Wahl von Bauelementwerten einige Beschränkungen auferlegt, besonders wenn zwei Spannungspegel benutzt werden.

Mit Magnetkopfwicklungen niedriger Induktivität ist diese Schalteranordnung einfach, zuverlässig und wirtschaftlich. Für Magnettrommeln mit Köpfen höherer Wicklungsinduktivität wird es erforderlich, Hochspannungstransistoren zu verwenden, was den Schalter verhältnismässig verteuert; die Einfachheit und Zuverlässigkeit der Anordnung bleibt jedoch erhalten.

Durch neue Schaltungsprinzipien verbesserte Thermoverzögerung von G. F. Klepp und J. E. Brooke-Stewart

Zusammenfassung des Beitrages auf Seite 106-109

Thermoverzögerungsschalter finden Anwendung für die Verzögerung im Einschalten der Anodenspannung von indirekt geheizten Röhren und Gleichrichtern, beim Anlaufen von Motoren, wiederholtem Schalten, sowie in zahlreichen anderen Verzögerungsschaltungen.

Thermoverzögerungsschalter haben zwei Hauptmerkmale: die Einschaltverzögerung und die Ausschaltverzögerung. Wenn die Endverzögerung nicht durch die Einschaltverzögerung allein, sondern durch Addition der Einschalt- und Ausschaltverzögerung erzielt wird, wird die Zeitstreuung zwischen den einzelnen Schaltern im Vergleich mit Anwendungsbeispielen, in denen nur eine der beiden Charakteristiken Verwendung findet, stark reduziert. Es wird gezeigt, dass die kombinierte Verzögerungszeit in hohem Masse von der Heizspannung und Umgebungstemperatur unabhängig ist und die Schalterlebensdauer geringen Schwankungen unterliegt.

Ein Komplementärpaar-Kaskodenverstärker für HF-Verstärkung von M. D. Wood

Zusammenfassung des Beitrages auf Seite 110-111

Man stimmt im allgemeinen überein, dass die Emitter-Basis-Kaskodenschaltung eine Alternative für die Neutralisierung darstellt. Dieser Beitrag vergleicht die Vorteile der Kaskode und der Neutralisierung und schlägt eine Kaskodenschaltung mit Komplementärpaar vor, dessen Vorteile über die herkömmliche Kaskodenschaltung—besonders bei Frequenzen über 10 . . 20 MHz—gezeigt werden.

Ein Gleichstromverstärker mit Spannungsverstärkungsfaktor 1 von G. G. Bloodworth

Zusammenfassung des Beitrages auf Seite 112-114

In einem entwickelten galvanisch gekoppelten Verstärker liegt der Spannungsverstärkungsfaktor ohne Versetzen zwischen Ein- und Ausgang sehr nahe bei 1; eine Aufhebung erster Ordnung der Drift von Transistorkenndaten mit Temperatur findet statt.

Gegenkopplung und Rauschkennwerte von R. E. Bogner

Zusammenfassung des Beitrages auf Seite 115-117

Es wird nach einem sehr einfachen Allgemeinverfahren festgestellt, dass die Rauschzahl und die optimale Quellenimpedanz von der Gegenkopplung unabhängig sind. Die Ableitung gibt Anregung zu einer bequemen Methode zur Bestimmung des Einflusses der in Gegenkopplungswegen benutzten Schaltung auf die Rauschkennwerte. Anwendungsbeispiele werden gegeben.