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Commentary

THE Institution of Electronic and Radio Engineers recently celebrated its fortieth birthday. The actual celebration took the form of a dinner at the Savoy Hotel, London, where the Guest of Honour was Admiral of the Fleet The Earl Mountbatten of Burma, who has been a member of the Institution for thirty years and has twice been its President.

As a 'Chartered Institution' the I.E.R.E. now takes its place with pride and dignity alongside the other learned societies of the world; but it has not always been so. In its early days it had a great struggle to survive and an even greater struggle to obtain recognition from other institutions and official organizations.

The formation of the Institution took place at a meeting of wireless engineers held in London on 31 October 1925. Among those present were Lord Gainford of Headlam, then Chairman of the British Broadcasting Company, and Sir William Noble, a Director of the British Broadcasting Company. The latter remained an active member of the Institution and a Vice-President until his death in 1943. It was intended, at first, that it should be called the British Institute of Radio Engineers but, to avoid any confusion with the American Institute of Radio Engineers, which had been founded in 1912, this was abandoned and it was agreed that the British body should be known as the 'Institute of Wireless Technology.' Although this title was regarded by many with disfavour, and in some ways it tended to hamper the early growth of the Institute, it was not until 1941 that the title was changed to the British Institution of Radio Engineers. The latter title was retained until 1964 when it was changed to the Institution of Electronic and Radio Engineers; a change which was made to more fully express the scope of its learned society activities.

From its inception in 1925 it was intended that the Institution should be one of high professional standards and that its membership should be composed of qualified engineers. To found such an Institution forty years ago was a brave, or foolhardy, undertaking; depending on one's point of view. In the early 1920s there were probably no more than one or two hundred engineers in this country who were interested in, or professionally qualified for membership of such an Institution; indeed, the membership list for 1926 shows a grand-total of 56! At that time there were very few organized courses of study of 'wireless' technology and thus the opportunities to qualify as a radio engineer were very limited; so too were the opportunities to practise as one since the output of the industry was confined to equipment for the Services, the General Post Office and the domestic entertainment market, and the demands of none of these was very large. However, despite the severe depression of the 1930s and the difficulties of the war years the membership grew steadily, if not spectacularly. By 1940 it had reached 1 425, by 1950 it had grown to 3 349 and by 1960 this had almost doubled to 6 332 while in February of this year the active membership topped the 10 000 mark.

Over the years the activities of the Institution have been those normal to a professional society. Considerable attention and encouragement has been given to the education and training of engineers; regular meetings have been held in London and in an increasing number of Centres in the U.K. and in countries overseas; representation has been supplied on various outside bodies and, by no means of least importance, a journal has been published. The first issue of the Journal appeared in October 1926 and the publication of the next few issues was somewhat sporadic. From 1932 to 1949 it oscillated between bi-monthly and quarterly publication. From 1949 onwards it has appeared as a regular monthly publication. A highlight of the Institution's activities has been the holding of a number of conventions which have attracted a worldwide attendance and it is interesting to note that the first of these, held at Blackpool in June 1934, was the first 'radio' convention to be organized by a European body and, so far as is known, was the first Television Convention ever to be held.

Over the years the Institution has received a number of honours and marks of recognition. High among these has been that of Royal Patronage. In 1946 His Majesty King George VI graciously granted His Patronage and this was followed in 1952 by that of Her Majesty Queen Elizabeth II. In 1951 His Royal Highness the Duke of Edinburgh accepted an invitation to become an Honorary Member. In 1955 the Institution made a successful petition for a Grant of Arms and in 1961 it achieved its greatest distinction by the award of a Royal Charter. This latter event completely changed the public image of the Institution and placed it on a par with the other great learned societies.

That the Institution of Electronic and Radio Engineers has reached its present position and status is due largely to the hard work and enthusiasm of a comparatively small band of men whose efforts, for the last 28 years of the Institution's history, have been ably marshalled by its General Secretary Mr. Graham D. Clifford, C.M.G. Their efforts have been well rewarded.

A Constant Volume Amplifier Covering a Wide Dynamic Range

By A. G. Morris*, B.Sc.(Eng.)

Advances in constant volume amplifier design have been limited by the need for an ideal variable gain device. The bottomed transistor is very close to the ideal as it produces very little distortion and its potential control range is in excess of 120dB. A system using the bottomed transistor is described which compresses an input variation of 60dB into 4dB. The utilization of this wide dynamic range, however, presents two further problems; first, the condition for optimum stability margin of a wide range control system must be established and second, additional circuits are necessary to prevent noise amplification between words. Solutions for these two problems are given, the solution for the former being of general application in control theory.

(Voir page 564 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 571)

IN speech processing systems it is usually desirable to prevent gross variations in signal level. In speech analysis applications, for which the circuit described in this article was developed, it is necessary for long term variations to be less than 10dB. This can be achieved either by physically controlling the speaker and the amplifier gain, or by controlling the amplifier gain automatically. The range to be covered is quite large, since in addition to variations of 30dB between different speakers, the speech of a single person during long conversations can vary by over 20dB. Also, since a handset is used with the equipment, large variations in signal can be caused by quite small head movements.

Very little data is available on the desirable characteristics for a constant volume amplifier, although it seems apparent that a gain variation of at least 60dB is required[1], and that the attack time should be as fast as possible. Also, although the gain must compensate for changes in input level, it must be held constant during silent periods in order to prevent the background noise from being amplified to the nominal output level. This last feature presents problems which can be very difficult to solve.

The Variable Gain Amplifier

The requirements for the amplifier were: wide dynamic range, low distortion, fast response time, and negligible thump. This last term is defined as the breakthrough of d.c. control voltage into the a.c. signal path. It should not be confused with noises caused by system instability, although the two usually interact.

Of the ten different methods of gain control examined, the only one that appeared to offer a complete theoretical solution was pulse width modulation. The signal would be chopped at a supersonic rate and amplified by an amplifier with a low-pass characteristic in order to eliminate the chopping frequency from the output. Varying the mark-to-space ratio of the chopping waveform by a factor of 1 000 to 1 would enable 60dB of gain variation to be obtained. The technical problems involved in producing and utilizing chopping pulses of the order of 25nsec caused this method to be dropped in favour of the bottomed transistor, which is very much easier to use. An additional advantage of the bottomed transistor over most of the other methods of gain control examined, is that its performance can be predicted with a fair degree of accuracy.

Varying the base current of a transistor can cause large variations in its saturation resistance while inducing only small collector voltages. A complete analysis of the low frequency performance of the transistor was described in a paper by Ebers and Moll[2] in 1954, and a brief resumé of this work is included here. The voltage-current relationship of an abrupt semiconductor junction can be described fairly accurately by the Shockley diode equation:

$$I = I_s \exp(qV/kT - 1) \quad \ldots\ldots\ldots\ldots \quad (1)$$

If this equation is applied to the two junctions of a common base transistor, equations can be derived which describe its performance accurately under all normal operating conditions. Current will flow across the emitter junction according to the equation, the number of current carriers reaching the collector junction being limited by the current gain α_n. At the same time, current carriers leave the collector junction, α_i of which reach the emitter. (The fact that only some of the current carriers cross the base region means that current will flow out of the base terminal.) By this means the emitter and collector currents can be specified. Thus:

$$I_e = I_{es}[\exp(qV_e/kT)-1]-\alpha_i I_{cs}[\exp(qV_c/kT)-1] \quad .. \quad (2)$$

$$I_c = I_{cs}[\exp(qV_c/kT)-1]-\alpha_n I_{es}[\exp(qV_e/kT)-1] \quad .. \quad (3)$$

After some manipulation of these expressions the emitter and collector potentials can be expressed explicitly, and the emitter-collector voltage is found to be:

$$V_{ce} = (kT/q)\,.\log_{\alpha_i}.\frac{I_b - I_c/\beta_n + I_{co}/\alpha_n}{I_b + I_c(1-\alpha_i) - I_{eo}} \quad \ldots\ldots\ldots\ldots \quad (4)$$

Differentiating this expression with respect to I_c gives the small signal bottomed impedance of the transistor:

$$R_B = (kT/qI_b)\,[(1/\beta_n) + (1/1+\beta_i)] \quad \ldots\ldots \quad (5)$$

If the base current is large, equation (4) becomes:

$$V_{ce} = (kT/q)\,.\log_{\alpha_i}$$

Symbols Used

I_e, I_c, I_b	=	emitter, collector and base currents
I_s, I_{es}, I_{cs}	=	diode, emitter and collector reverse saturation currents
I_{eo}, I_{co}	=	emitter and collector leakage currents
V_e, V_c	=	emitter and collector potentials relative to base
V_{ce}	=	collector-emitter potential
kT/q	=	25mV
α_n, α_i	=	normal and inverted current gains common base
β_n, β_i	=	normal and inverted current gains common emitter

* *The M.E.L. Equipment Co. Ltd.*

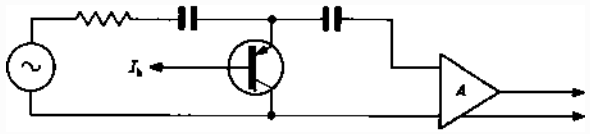


Fig. 1. Transistor operated as a shunt attenuator

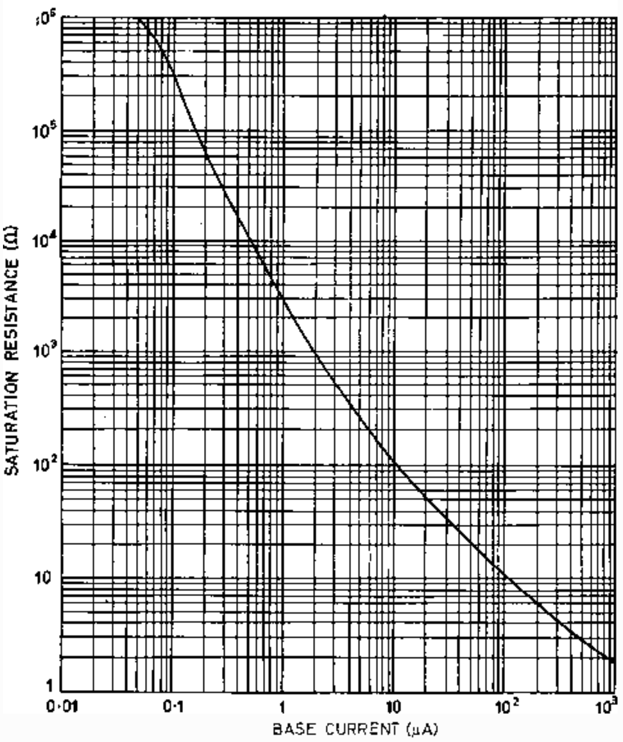
$$\therefore V_{ce} = \frac{kT}{q(1+\beta_i)} \quad \dots\dots\dots\dots\dots\dots \ (6)$$

if β_i is large.

Although the algebra involved is rather tedious, the final equations are simple and very useful. If a transistor is operated as shown in Fig. 1 it provides an almost ideal attenuator. As can be seen from equation (5), saturation resistance is inversely proportional to the base current, and the voltage induced at the floating electrode due to the base current is generally less than 1mV. In practice, the minimum resistance is limited by the resistivities of the emitter and collector regions, and alloy transistors are to be recommended in this respect. Also not included in equation (5) is the effect of leakage currents which limits the maximum impedance to a few megohms.

A curve of bottomed impedance for a typical transistor, type OC201, is shown in Fig. 2. It can be seen that, apart from the above mentioned limits, there is a distinct curvature of the characteristic due to the variation of current gain with emitter current. A small change in α_n can result in a large change in β_n. This does not apply to the inverse current gains, since β_i is usually between 2 and 3 for the OC201. As can be seen from equation (5), since R_B depends mainly on the smaller of the two current gains, the curve can be expected to be fairly stable with time for a particular sample and not to vary greatly from sample to sample of a given type. This fact becomes important when considering the open loop response of the complete system. The advantage of operating the transistor in the inverted mode is well known; the larger of the two current gains can be used in equation (6), resulting in a very small induced collector voltage which is independent of base current.

Fig. 2. Transistor saturation resistance



The only remaining parameter of interest is the distortion produced by the attenuator. This can be determined by assuming a sine function for the collector current and expanding equation (4). If the log series is used, the algebra becomes very tedious since each coefficient is a series itself. An alternative method is to assume a Fourier series and evaluate the coefficients. This results in integrals of the type:

$$a_n = (1/\pi)\int_{-\pi}^{\pi} \cos n\Theta \,.\, \log(1+K\,.\,\sin\Theta)\,.\,d\Theta \quad \dots\dots\dots \ (7)$$

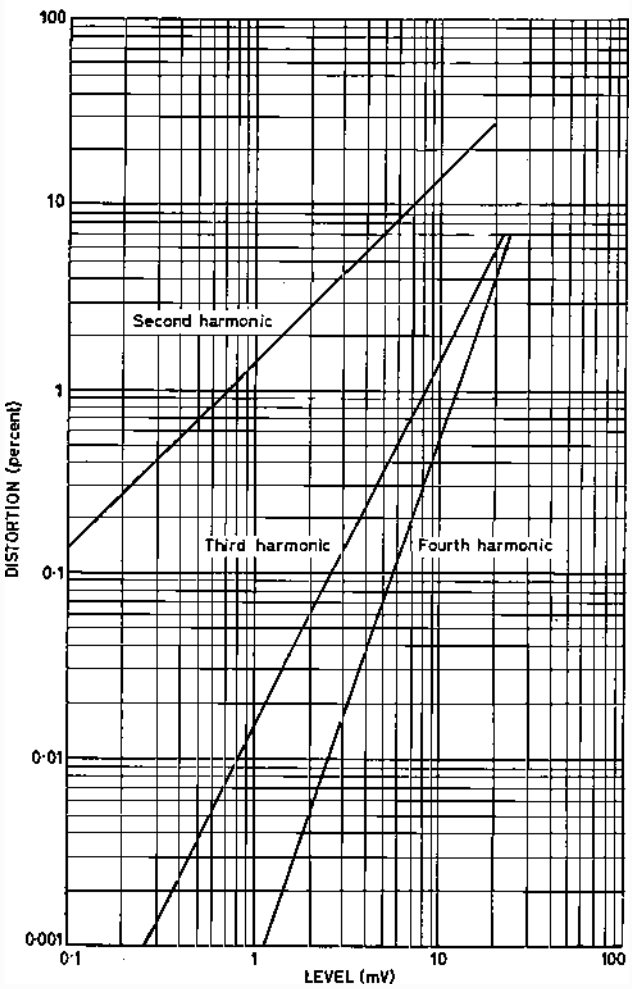
For the evaluation of this integral, the author is indebted to Professor A. Stone[3] of Rochester University. Space does not allow the inclusion of more than the result of his elegant derivation which yields:

$$\log(1+K\,.\,\sin\Theta) = \log[1+m^2]/m + 2[m\,.\,\sin\Theta+(m^2/2)\cos 2\Theta-(m^3/3)\sin 3\Theta-(m^4/4)\cos 4\Theta+\,..] \quad \dots\dots\dots\dots \ (8)$$

where $m = (1/K)(1-K^2)^{\frac{1}{2}}$

This can be substituted in equation (4) and the distortion calculated for different signal levels. Fig. 3 shows the calculated distortion for a transistor in which $\beta_n \gg \beta_i$. A very interesting point comes out of the calculation of these curves: the distortion depends only on the signal level and is independent of transistor parameters. It follows

Fig. 3. Theoretical attenuator distortion



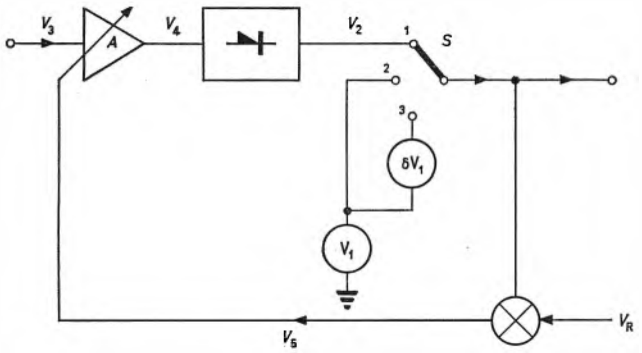


Fig. 4. Arrangement for derivation of loop gain

from this that distortion produced is independent of operating point or type of transistor (providing it is asymmetric). Distortion produced by symmetrical transistors is marked by the absence of even harmonics, the odd harmonics having the same values as shown as Fig. 3. Measured distortion for the OC201 agrees fairly well, being about 15 per cent lower than the calculated values.

Stability

It is generally realized that the main problem with feedback systems is the question of stability. What is not generally appreciated is that there are two basically different types of feedback system and that the stability problems of each need entirely different approaches. First, there is the feedback amplifier in which a fraction of the output is added to the input at the input of the amplifier. The gain will tend towards the reciprocal of the feedback factor. The characteristics of such a system are output proportional to input, and constant loop gain (at one frequency). The second type of system, which includes servos and regulators, compares the output of the amplifier with a reference, and any difference is fed back to adjust the gain. The output will tend towards a reference level and be largely independent of input level. The forward gain will, of course, be approximately inversely proportional to the input level. Unfortunately, the loop gain can also follow the same law, resulting in the system stability changing with input level.

The normal practice with feedback systems is to ensure that there is only one major phase shift in the loop. By this means, at high frequencies, where minor phase shifts become significant, the dominant phase shift will have caused the loop gain to drop to less than unity. However, if the loop gain varies with input level, and the input can vary over a range of 60dB, this method cannot be expected to be sufficient to keep the system stable. On the other hand, if the loop gain is low, the required output regulation will not be achieved.

It is therefore a considerable advantage if the loop gain can be made constant and independent of input level. The loop gain can then be increased until the system approaches instability for a given input. The margin of stability will then be the same for all inputs and the regulation will have its best practical value. The requirements for such a controller will now be established.

Fig. 4 shows a feedback system with an input V_3, and an output V_4 rectified as V_2. It is convenient for this analysis to define the gain as:

$$A = V_2/V_3 \quad \text{.............. (9)}$$

V_R is the reference voltage and is subtracted from V_2 to yield the error voltage V_5. If S is switched to position 2, V_1 can be adjusted to give the previous output.

Thus:

$$V_1 = V_2 \quad \text{.............. (10)}$$

Switching to position 3 causes all the variables in the loop to be changed by small amounts. Using primed symbols for their new values,

$$V'_1 = V_1 + \delta V_1$$
$$A' = A + \delta A$$
$$= (V_2/\delta V_1) \cdot (\delta A/A)$$
$$\therefore \delta V_2 = V_2 \cdot (\delta A/A) \quad \text{.......... (11)}$$

loop gain, $L = \delta V_2/\delta V_1$

$$= (V_2/\delta V_1) \cdot (\delta A/A)$$

Substituting equation (10) into this,

$$L = (\delta A/\delta V_1) \cdot (V_1/A) \quad \text{.......... (12)}$$

It is now possible to synthesize the controller characteristic required for constant loop gain. Re-arranging and integrating equation (12),

$$L \cdot \log V_1 = \log (A/A_0) \quad \text{........... (13)}$$

where $\log A_0$ is the integration constant.

i.e. the required condition is that $\log A$ should vary linearly with respect to $\log V_1$. However, since the system regulation is good, it is reasonably accurate to expand $\log V_1$ as $V_5/V_R + \log V_R$. Under these conditions equation (13) becomes:

$$\log A = k_1 \cdot V_5 + k_2 \quad \text{............ (14)}$$

This solution is more easily realizable in practice since only one log function is required. It results, however, in a departure from the ideal of constant loop gain. The loop gain of a system obeying equation (14) can be evaluated by applying equation (12). Hence:

$$L = k_1 \cdot V_2 \quad \text{.............. (15)}$$

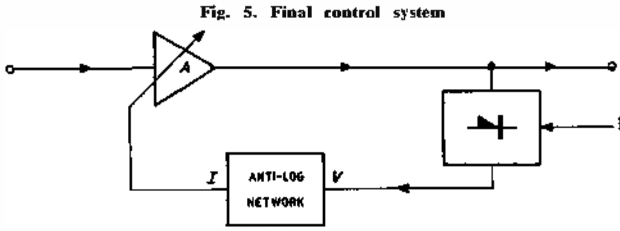
i.e. the loop gain will vary in proportion to the output voltage. This will not cause any difficulty since the higher the loop gain the more constant will the output voltage be. The suggested arrangement is shown in Fig. 5. Silicon diodes are used for the logarithmic convertor, with the result that the base current of the gain control transistor is logarithmically related to the error voltage.

It is interesting to consider a system in which the control voltage is applied to the gain control transistor via a resistor. The gain relationship will then be linear, i.e.

$$1/A = a \cdot V_5 + b$$
$$\therefore \quad L = -a \cdot V_5 \cdot A$$

If such a system is to cope with an input variation of 60dB, the loop gain will vary by the same amount. Thus either instability, long attack time, or poor regulation must result. In practice it is found that an attack time of some 5sec is necessary in order to make the system stable over the entire input range, although instability is still observed for small inputs. If the input is abruptly increased by 10dB the output will momentarily increase by this amount and then vanish completely due to the application of excess control around the loop. This phenomenon has also been described by Pope[4], although the observed instability was

Fig. 5. Final control system



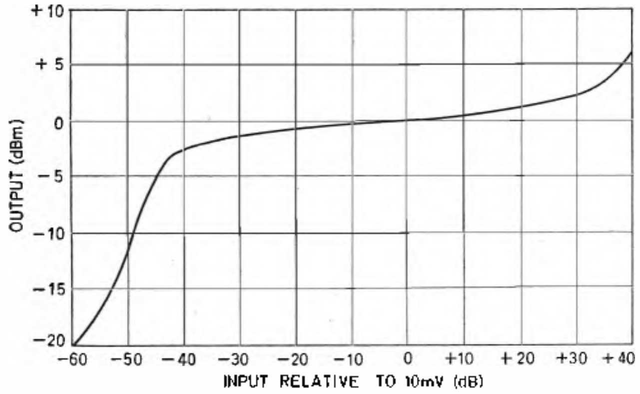


Fig. 6. Regulation curve

probably enhanced by the large thump voltages produced by the base-emitter diode control circuit used.

Another example of this type of problem is the thermostatically controlled oven. The amount of heat required for a 1°C rise at low temperatures is much smaller than that at high temperatures. Thus, at low temperatures, the loop gain is very high and large amplitude oscillations occur. A non-linear controller is essential if these oscillations are to be eliminated without sacrificing either response time or temperature range.

The regulation curve of the constant volume amplifier is shown in Fig. 6. It is linear over the central 60dB, and it is therefore very convenient to monitor the input level by connecting a meter to the control voltage. This meter can be calibrated directly in decibels, the graduations being equally spaced over the 60dB of input variation. A further advantage of the logarithmic control voltage is that a constant current discharge of the control voltage storage capacitor results in the rate of gain recovery being the same at all points on the regulation curve.

Noise Suppression and Speech Detection

A gain recovery rate of 10dB/sec is found to be suitable for fairly rapid speech and is capable of following changes in level due to most causes. However, during pauses in the speech the stored control voltage will decay, allowing the background noise of the input signal to be amplified to the nominal output level. Since the noise level of speech is frequently only 30dB below the signal, only 3sec need elapse before the noise at the receiving end becomes intolerable and two-way speech impossible. Decreasing the recovery rate to 1dB/sec allows noise-free pauses in the speech to take place but does not offer a long-term solution.

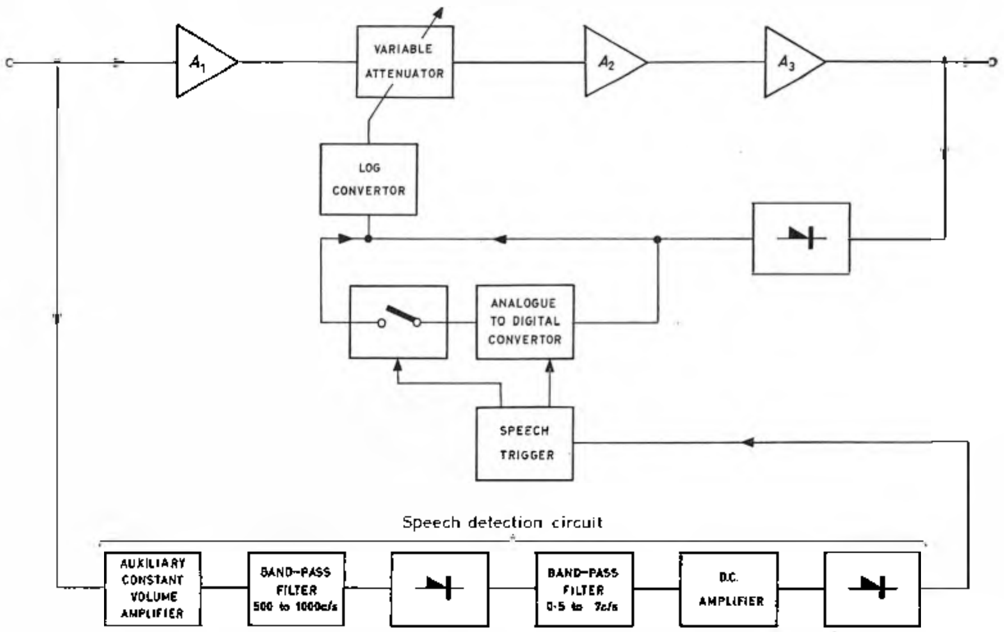
The method of noise suppression finally adopted has two rates of gain recovery: 10dB/sec during speech and 0dB/sec at other times. Thus, during silent periods the gain is held constant at its previous value and only allowed to change when speech recommences. The control voltage can be held in a capacitor-buffer amplifier store, but it was decided that protracted measurements would be difficult and a simple analogue to digital convertor was therefore designed. Even so, attention is drawn to the fact that the buffer amplifier must have an input impedance of at least 1 000MΩ if it is to cope with normal two-way conversations.

The detection of speech cessation can be accomplished by feeding the control voltage into a trigger circuit. This is fairly successful although the speaker may occasionally recommence speaking at a substantially lower level. Under these conditions the system would act as a fixed gain system and the output would be barely audible. It is obvious that a more sophisticated method of speech detection is required, one which is completely independent of the main amplifier. The method eventually used consists of an auxiliary constant volume amplifier followed by a syllabic modulation detection circuit similar in principle to one described by Getgen[5]. The complete block diagram is shown in Fig. 7. The auxiliary constant volume amplifier amplifies all inputs to the nominal output level. This is followed by a 500c/s to 1kc/s band-pass filter which separates the first speech formant. The output of this filter is rectified in order to allow the examination of the envelope of the speech input. It has been established that speech syllables last from about 2sec to about 150msec for stop consonants[6]. The rectified formant is therefore passed through a filter which has a pass-band extending from 0·5c/s to 7c/s. The rectified output of this filter can now be used to operate the speech trigger. This trigger circuit will thus only operate when a very specialized signal is used, one which consists of the first speech formant modulated at the normal syllabic rate.

Circuit Description

The circuit of the constant volume amplifier is shown in Fig. 8, with auxiliary circuits shown in block form. Transistors VT_1 and VT_2 form a buffer stage and feed a constant current signal into attenuator transistor VT_3. The attenuation produced is therefore proportional to the control current. Transistors VT_4 to VT_9 amplify the signal to about 3V r.m.s. 60dB of gain is required and this is

Fig. 7. Arrangement of the equipment



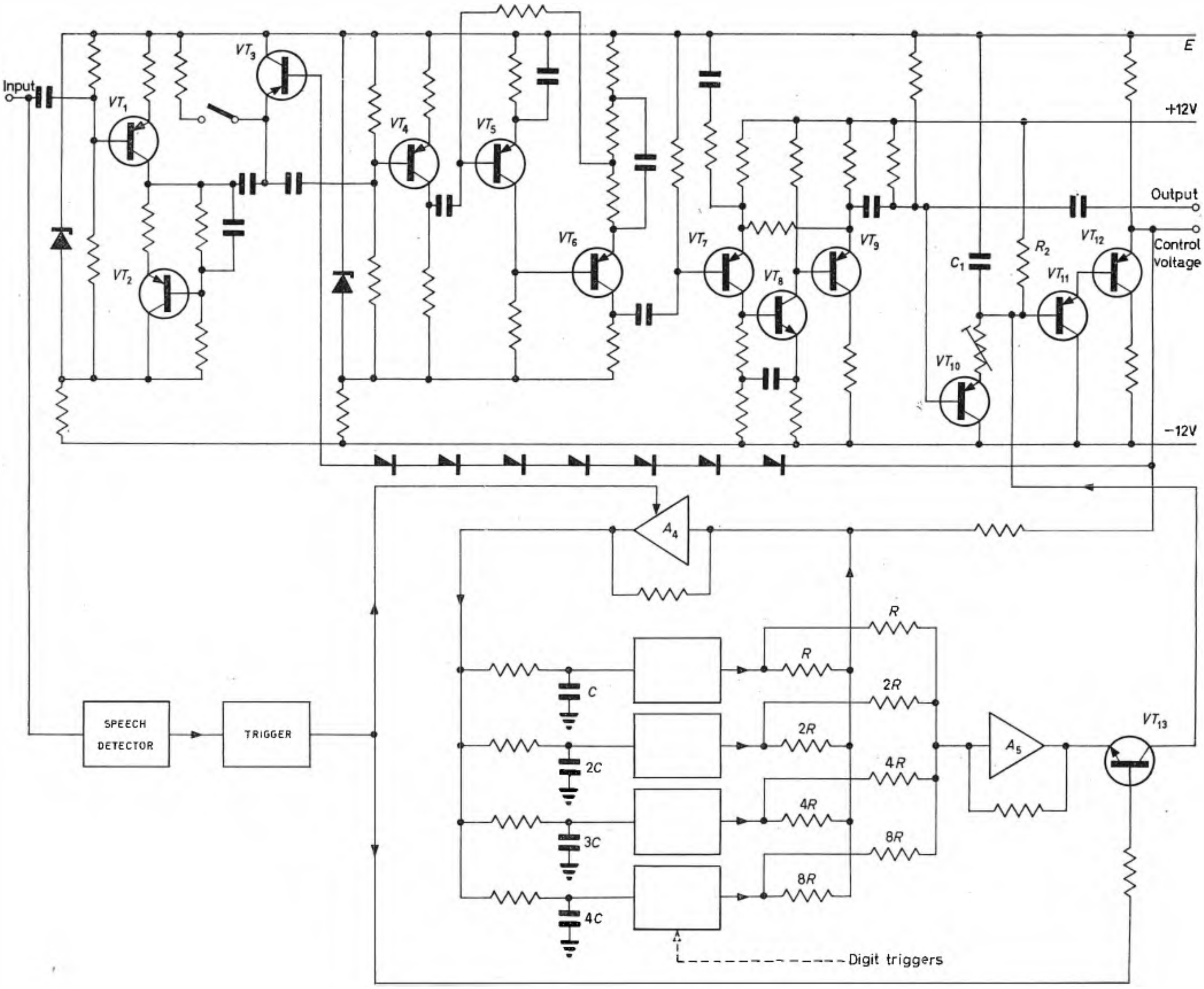


Fig. 8. The constant volume amplifier circuit

supplied by two amplifiers, each having 30dB of gain and 30dB of feedback in order to ensure high gain stability. Although at first sight it appears that the output will be unaffected by changes in amplifier gain, the only thing that remains unaltered will be the output amplitude. If the gain is too low, the signal across VT_8 will be large enough to cause significant distortion. If the gain is too high, circuit noise will be amplified excessively.

The signal output appears at the emitter of VT_9 and is rectified by buffer-rectifier VT_{10}. The reference voltage is supplied to VT_{10} by two resistors across the supply. A highly accurate reference is not necessary since a change in reference of 10 per cent will change the nominal output by only 1dB. It should be noted that to this reference must be added the voltage drop across the diodes in the logarithmic convertor.

The attack time can be varied by adjusting a resistor in series with the rectifier. The discharge rate of the reservoir capacitor C_1 is controlled by R_2 which is returned to +12V, resulting in an approximate constant current discharge of the negative control voltage. The control voltage is buffered by a compound emitter-follower and then fed to the logarithmic diodes and also the analogue to digital convertor. The diodes used are type OA200 which give an accurate logarithmic conversion from 0·2μA to 3mA. During speech the analogue to digital convertor continually digitizes and the output of amplifier A_5 continually follows the control voltage (to within one bit). The convertor consists of four trigger circuits with a backlash of 2V, and a comparator amplifier A_4. The outputs of the four digit triggers are fed to the comparator via four weighting resistors. The control voltage is also fed in at this point via a fifth weighting resistor.

The output of amplifier A_4 is used as the input to the trigger circuits, each of which is supplied via a different CR delay circuit. These delays are arranged so that the least significant digits will be operated first. There are four more weighting resistors connecting the digit triggers to a second summing amplifier A_5 which thus develops a continuous copy of the control voltage to within one bit (200mV). When speech stops, the comparator voltage is frozen to the half backlash point, and the digital voltage is used for gain control.

The speech detection circuit is shown in more detail in Fig. 7. The auxiliary constant volume amplifier is a simplified version of the main amplifier since the requirements are not so stringent. Both band-pass filters are essentially the same, consisting of an active low-pass filter with a third order Butterworth response[7] followed by a similar high-pass filter. The output from the second filter is rather low and is amplified by a simple d.c. amplifier having push-pull outputs to improve the speed of rectification at low frequencies.

Performance of Circuit

The regulation curve of the circuit is shown in Fig. 6. For an input variation of 60dB the output varies by 4dB. This curve was plotted using a sine wave input, with the speech detection circuit biased permanently on. As expected, the distortion was 3 per cent and did not vary much over the control range. In practice, since the r.m.s. value of speech is normally about 10dB below the peak level, the effective distortion is nearer 1 per cent. A higher degree of compression is possible, but only at the expense of fast attack time, since both the signal and the supply voltages are generally accompanied by large narrow spikes which can only be ignored if either the loop gain is low or the rectifier has a low-pass characteristic.

The ideal of a very fast attack time proved in practice to be rather misleading. If the rectifier series resistance is set to zero (fast attack time), with normal input, there is a curious grabbing effect as if the speaker were taking a violent inward breath without interrupting his flow of speech. This phenomenon is due to the fact that certain sounds such as plosives are very much louder than other sounds, and if control is applied so that their amplitude is equal to the nominal output level, subsequent sounds will be muted. If this were carried to its conclusion, all variation in voice intensity would be eliminated and a great deal of the voice intonation would be destroyed.

An attack time of 10msec was found suitable, although a longer attack time will not have much effect on the speech output. It was not found practical, however, since the decreased loop gain had an adverse effect on the regulation. The value of series rectifier resistance used was 200Ω, and if two silicon diodes are connected in parallel with this the circuit will be able to react quickly to gross overloads.

Tests on the speech detection circuit using normal speech resulted in speech being indicated for about 30 per cent of the time. Taking into account the pauses between words, this is surprisingly good. It is, of course, more than sufficient for adequate control, and to allow the system to follow voices which rapidly change in volume. During quiet periods the gain was held constant. Tests using random noise, teleprinter signals and other sounds did not result in a single faulty indication, indeed the output from the syllabic filter was observed to be very small.

The amplifier is set up by switching to the fixed mid gain position, and the attenuator between the microphone and constant volume amplifier adjusted to give the nominal output level for normal speech input. The switch is then moved to the automatic position and the system will control itself for all speech levels within the dynamic range. However, if the power supply is switched off, and then switched on again, the digital store can take up any value. This may result in quiescent background noise being amplified to the nominal output level until speech commences.

Acknowledgments

The work reported here was carried out for the Signals Research and Development Establishment. The author wishes to thank S.R.D.E. and the Directors of The M.E.L. Equipment Company Ltd. for permission to publish this article. Thanks are also due to the author's colleagues, Mr. J. E. Brown and Mr. J. N. Evison, whose suggestions helped to solve the many problems encountered.

REFERENCES

1. Purton, R. F. A Survey of Speech Signal Statistics and their Significance in the Choice of a P.C.M. Companding Law. *Proc. Instn. Elect. Engrs. 109B*, 60 (1962).
2. Ebers, J. J., Moll, J. L. Large-Signal Behaviour of Junction Transistors. *Proc. Inst. Radio Engrs. 42*, 1761 (1954).
3. Stone, A. Private communication from the University of Rochester, New York (1962).
4. Pope, G. E. Transistor Constant Volume Amplifier. *Wireless World. 66*, 88 (1960).
5. Getgen, L. E. Amplifier Compensates for Speech Level Variations. *Electronics. 33*, 103 (29 July 1960).
6. House, A. S. On Vowel Duration in English. *J. Acoustical Soc. Amer. 33*, 82 (1961).
7. Bach, R. E. Selecting R.C. Values for Active Filters. *Electronics. 33*, 82 (13 May 1960).

Military Communications Equipment

The Plessey A13 radio is a versatile, multi-purpose set, available as a light-weight one-man pack at its simplest to a high power, long range base radio station.

It is the first fully-transistorized equipment of its type to be introduced for use by the British Army. Developed by Plessey in conjunction with the Signals Research and Development Establishment of the British Ministry of Aviation, it meets the British Government Specification K114 and, as a result of extensive user trials, has been officially accepted by the British Armed Forces and is now in full production.

The equipment is already in service with both British and Commonwealth armed forces and has proved entirely satisfactory over the tough terrain of the Middle and Far East.

The equipment provides phase modulated, amplitude modulated and c.w. modes of operation. Phase modulation was chosen as the primary mode for this equipment because of advantages over amplitude modulation in range and power consumption. It also allows the set to remain simple and light. In addition to its small bulk, low weight and low battery drain, a significant feature of the equipment is its exceptional frequency stability which allows for any of the 2400 channels in the 2 to 8Mc/s frequency range to be selected and tuned without netting.

The basic unit, a transistorized transmitter/receiver in a light but tough die-cast case, can be adapted by the addition of ancillary units for a wide range of applications. Output power, for example, of the basic set is 1·5W, whereas the high power version gives 16W.

Maximum operating range for minimum battery consumption is achieved by using phase modulation. This mode, compared with a.m., effects up to 30 per cent better ground wave coverage for the same battery consumption. Additional advantages are increased operating time and a reduction of overall station weight. To allow for inter-operation with existing systems, provision for amplitude modulation is included and c.w. working up to 25 words/min is possible.

The receiver is easily tuned to any of the 2400 channels in the band 2 to 8Mc/s without netting, and a meter on the front panel makes tuning simple. No crystal bank is required and the set is exceptionally stable over a wide range of temperatures. The calibrator provides check points at 100kc/s and 10kc/s throughout the band. After selecting the wanted frequency range the operator calibrates the scale to 100kc/s and then to 10kc/s. Final tuning to 2·5kc/s is effected by moving the scale so that the required frequency coincides with the cursor position, an operation which is quickly mastered.

Throughout the development programme the closest liaison was maintained with the Electrical Inspection Directorate of the Ministry of Aviation and the R.E.M.E. Maintenance Advisory Group, and the final equipment incorporates their recommendations. The basic unit is built on modular lines, using printed circuit boards extensively, and maintenance can be carried out quickly and easily by semi-skilled personnel on a substitution basis. The boards are designed to survive rough handling during maintenance.

The operating temperature range is −32°C to 55°C (storage −40°C to 70°C). The equipment can be operated without retuning for at least one hour during which time the ambient temperature may change by up to 30°C.

Microwave Tunnel-Diode Oscillators

By J. Swift*, B.Sc.(Melb.), M.I.R.E.E.(Aust.)

This article discusses the application of tunnel diodes in microwave oscillators. Following an introduction, the parameters of the tunnel diode are discussed followed by its application in microwave oscillator circuits. Finally, full design information for a 2Gc/s oscillator is given and a bibliography of the work in this field is provided.

(Voir page 564 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 571)

THE first paper[1] on tunnel diodes appeared as recently as 1958. Since then the tunnel diode has found increasing application in amplifiers, oscillators, convertors, switching circuits and other devices[2,3]. The tunnel diode has a heavily doped pn junction that exhibits a dynamic negative resistance at a small forward bias. The tunnel diode derives its name from the phenomena of quantum mechanical tunnelling[1,2]. Most diodes are made of gallium arsenide or germanium. Transit time effects do not limit the performance in the microwave region and the negative resistance remains effective to very high frequencies, hence they have important microwave applications.

The present article is confined to the use of tunnel diodes in microwave oscillators. At the present time, commercially available diodes are limited to a frequency of about 15Gc/s but experimental diodes have been made that operate at frequencies exceeding 100Gc/s. Typical oscillator power outputs range from 10mW at 0·6Gc/s to 0·01mW at 7Gc/s.

Tunnel Diode Parameters

THE STATIC CURRENT-VOLTAGE CHARACTERISTIC

A typical static current-voltage characteristic curve for a tunnel diode is shown in Fig. 1.

The curve *ab* in Fig. 1 represents the region where the resistance of the diode is negative. The value of interest is not the total value of the diode resistance but the incremental value given by $R_j = dV_d/dI_d$. The peak forward current I_p is the value at which the slope of the characteristic curve changes from positive to negative with V_d increasing. For a given diode, I_p depends mainly on the resistivity of the crystal and on the junction area, and ranges from about 10 to 50mA for typical microwave tunnel diodes. The valley point current I_v is the value at which the slope of the characteristic curve changes from negative to positive with V_d increasing. The peak to valley current ratio I_p/I_v determines the current swing of the device and is about 10 for a typical gallium-arsenide tunnel diode. The peak voltage V_p and the valley voltage V_v are the values corresponding to the currents I_p and I_v respectively. V_v is normally only about 500mV and for this reason the operating bias and hence the power output of a tunnel diode are small. The forward voltage V_f is the voltage at which the current reaches the maximum peak current of the diode. The parameters I_p, I_p/I_v, V_p, V_v and V_f are provided by the manufacturer and enable a portion of the performance of the tunnel diode to be calculated.

THE EQUIVALENT CIRCUIT OF A TUNNEL DIODE

The encapsulated tunnel diode can be represented approximately to very high frequencies by the equivalent circuit shown in Fig. 2.

R_s is produced mainly by the resistance of the contact, base and external leads and is typically 1 to 2Ω. L_s is mainly the inductance of the mount, while C_j is the junction capacitance, which is a function of the applied voltage and is usually specified at the valley voltage V_v. Since C_j may be as high as 100pF in microwave tunnel diodes L_s must be kept low. By fabricat-

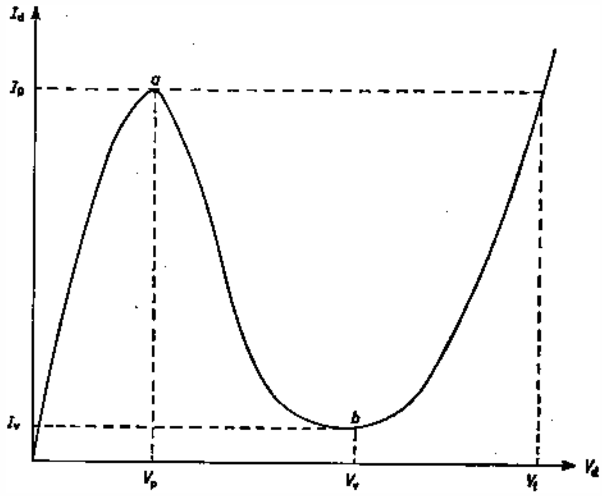


Fig. 1. Static current-voltage tunnel diode characteristic

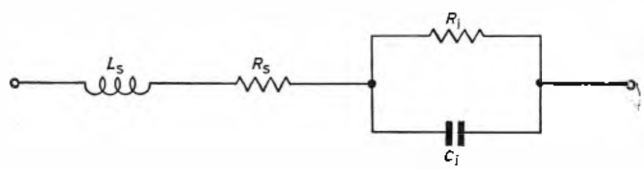


Fig. 2. Approximate equivalent circuit of a tunnel diode

ing the diode so that the mount forms a continuation of the external circuit, values of L_s of the order of 1nH are obtained. The junction resistance R_j which is a function of voltage and is negative may vary from 4 to 100Ω in commercially available diodes.

The impedance of the encapsulated diode (Z_d) is given by:

$$Z_d = R_s - \frac{|R_j|}{1+\omega^2 C_j^2 R_j^2} + j\left[\omega L_s - \frac{\omega C_j R_j^2}{1+\omega^2 C_j^2 R_j^2}\right] \qquad \ldots\ldots\ldots\ldots (1)$$

The corresponding expression for the admittance (Y_d) is most unwieldy and when Y_d is required it is better to calculate Z_d from equation (1) and find the reciprocal. The impedance (Z_d) is plotted in Fig. 3 for the gallium arsenide diode 40058.

If the real part of equation (1) vanishes:

$$R_s = \frac{|R_j|}{1+\omega^2 C_j^2 R_j^2}$$

$$\therefore\ f_{ro} = \frac{1}{2\pi |R_j| C_j} \cdot \sqrt{((|R_j|/R_s) - 1)} \qquad \ldots\ldots\ldots (2)$$

* The University of New South Wales, Australia.

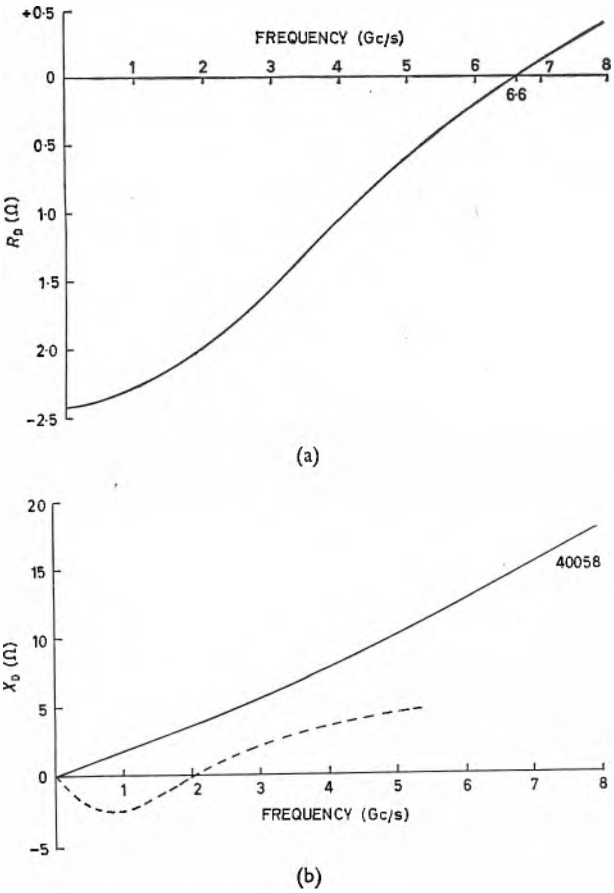


Fig. 3(a). R_D versus frequency for the RCA 40058 tunnel diode
(b). X_D versus frequency for the RCA 40058 tunnel diode

f_{ro} is called the resistive cut-out frequency and above f_{ro} the real part of Z_d is positive. Thus f_{ro} sets a limit to the operation of the tunnel diode since once the real part of the impedance ceases to be negative, oscillation cannot occur. For the 40058 gallium arsenide tunnel diode in which $|R_j| = 4{\cdot}4\Omega$, $R_s = 2{\cdot}0\Omega$ and $C_j = 6{\cdot}0$pF, $f_{ro} = 6{\cdot}6$Gc/s. This is shown in Fig. 3 as the point where the resistive portion of Z_d is zero.

If the reactive portion of Z_d in equation (1) vanishes:

$$\omega L_s = \frac{\omega C_j R_j^2}{1 + \omega^2 C_j^2 R_j^2}$$

$$\therefore f_o = \frac{1}{2\pi |R_j| C_j} \cdot \sqrt{((R_j^2 C_j / L_s) - 1)} \quad \ldots\ldots\ldots\ldots (3)$$

f_o is the self-resonant frequency of the tunnel diode. Below f_o the reactive part of Z_d will be capacitive and above f_o it will be inductive. However some tunnel diodes have no self-resonant frequency. If $R_j^2 C_j / L_s > 1$ there is a self-resonant frequency, but if $R_j^2 C_j / L_s \leqslant 1$ the term in the square root sign of equation (3) is zero or negative and f_o does not exist.

In Fig. 3 it is seen that the 40058 tunnel diode has no self-resonant frequency, the reactive part of Z_d being inductive at all frequencies. The behaviour of the reactive portion of other diodes with a self-resonant frequency is indicated by the dashed curve in Fig. 3.

Tunnel Diode Microwave Oscillators

A Basic Oscillator Circuit

A tunnel diode will oscillate whenever an operating point is established in the negative resistance region of the diode characteristic and a suitable resonant circuit exists. A basic tunnel diode oscillator circuit for operation below the self-resonant frequency of the diode is shown in Fig. 4.

Since the diode presents a capacitive susceptance at a frequency below self-resonance, an inductance L is provided. The resistance R is placed close to the diode and the inductance of the bias leads L_b does not affect the performance if R_1 and R_2 are much larger than R.

R serves three purposes:

(1) From a d.c. bias point of view R must provide a monostable operating point on the negative resistance portion of the diode characteristic. This is illustrated in Fig. 5.

The correct value for R is given by:

$$R_s + R < |R_j| \quad \ldots\ldots\ldots\ldots (4)$$

In this case the loadline which has a slope of $1/R$ will cut the diode characteristic at only one point P and the desired monostable operation will be obtained.

(2) R may serve as the r.f. load. To start oscillation the real part of the total circuit admittance (diode and external circuit) must be negative. Thus for oscillation:

$$\frac{R}{R^2 + \omega^2 L^2} < |g_d| \quad \ldots\ldots\ldots\ldots (5)$$

where g_d is the conductance of the encapsulated diode. This corresponds to the r.f. loadline cutting the diode characteristic in more than one point as shown by the dashed line in Fig. 5.

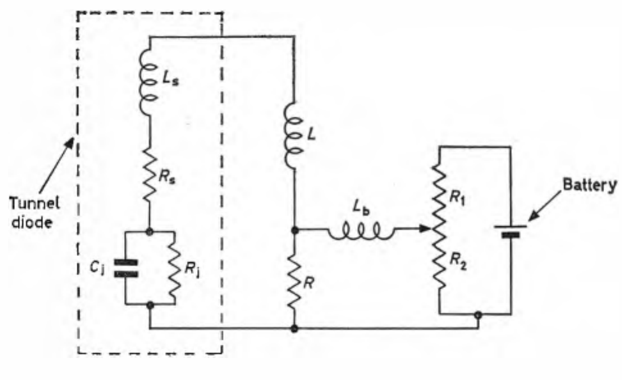
(3) R serves as a stabilizing or damping resistor to prevent low-frequency parasitic oscillations that can occur via the bias circuit leads. If equation (4) is satisfied then viewed from the bias supply the input admittance is positive and no low-frequency parasitic oscillations can occur.

As will be shown, the microwave load can be provided independently of R which then serves to provide a monostable operating bias and to suppress parasitic oscillations.

A basic oscillator circuit for use at frequencies where the diode is above the self-resonant frequency (i.e. the diode is inductive) is shown in Fig. 6. This circuit is also used when the diode has no self-resonant frequency and always presents an inductive susceptance.

The oscillator circuits shown in Figs. 4 and 6 are the simplest from an analytical point of view but the calculation of the power output and frequency requires the solution of non-linear differential equations. This has been

Fig. 4. A basic tunnel diode oscillator ($f < f_o$)



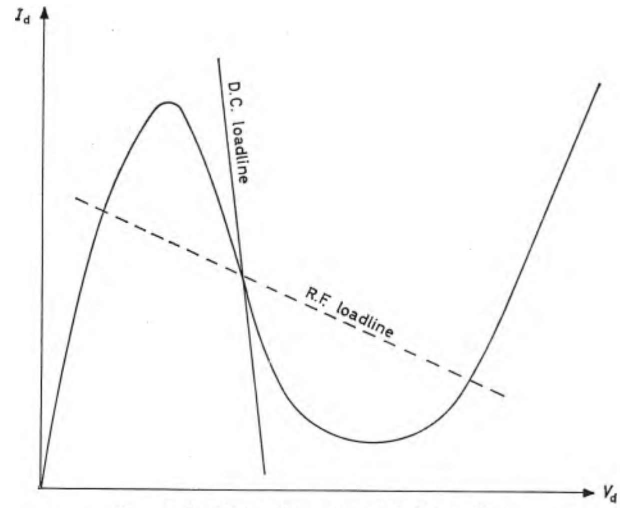


Fig. 5. Load lines for a tunnel-diode oscillator

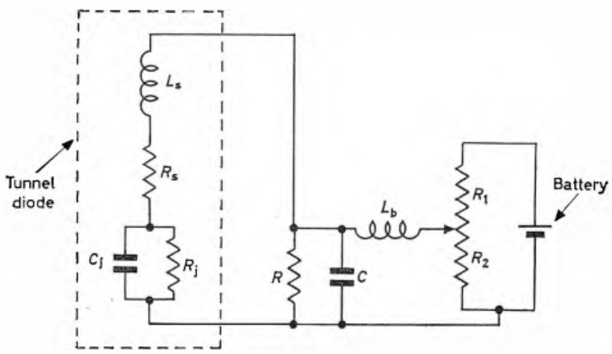


Fig. 6. A basic tunnel diode oscillator ($f > f_0$)

done in specific cases using a computer, but a closed form general solution is not possible[4].

An approximate expression for the power output of the tunnel diode (P) is given by[4]:

$$P \simeq 3/16\ (V_v - V_p)\,(I_p - I_v) \qquad (6)$$

Good agreement with experimental results has been obtained with equation (6) when $R_s \ll R_j$ and the operating frequency is not more than about one-third of the resistive cut-off frequency (f_{ro}).

A discussion of the stability of tunnel diode circuits has been given by several authors[4,5,6] and will not be included here.

Microwave Oscillators

When tunnel diode oscillators are operated at frequencies above a few hundred megacycles per second, coaxial lines, strip transmission lines, ridge waveguide and cylindrical cavities may be used as resonators. In these cases, the design is greatly simplified if it is assumed that the tunnel diode and the stabilizing resistor are lumped elements connected to distributed, uniform transmission lines. Tunnel diodes and microwave stabilizing resistors are so small that this assumption is usually valid to at least 10Gc/s.

The bibliography at the end of this article contains most of the references to the present time on the various microwave tunnel diode oscillators that have been built.

A simple and versatile oscillator circuit suitable for frequencies up to about 2Gc/s is shown in Fig. 7.

The resonator consists of an open-ended microstrip line[7] with the tunnel diode clamped to one end and output coupling and capacitive tuning at the other. Microstrip is most convenient because of the small physical size of the diode. The microwave stabilizing resistor (R) is placed one-quarter wavelength from the open end of the line and is therefore at a r.f. voltage null. The line length l is chosen to present an equal and opposite susceptance to that of the diode at the operating frequency. The output probe is adjusted to give the optimum load conductance to the diode and the frequency may be lowered by increasing the capacitive loading at the end of the line. In this circuit the resistor R is isolated from the microwave load because it is at a null of r.f. voltage and serves to provide a monostable bias point and suppress parasitic low-frequency oscillation as already discussed. In addition, little microwave energy is dissipated in the resistor over a reasonable frequency range, in contrast to other commonly used circuits[4].

Design of a 2Gc/s Oscillator

The design is based upon the oscillator of Fig. 7 and employs the RCA gallium arsenide tunnel diode 40058. Data on this diode is given in Table 1.

TABLE 1
40058 Tunnel Diode

$I_p = 50$mA	$L_s = 0{\cdot}4$nH
$I_p/I_v = 12$	$R_s = 2{\cdot}0\Omega$
$V_p = 180$mV	$C_j = 6$pF
$V_v = 570$mV	$R_j = -4{\cdot}4\Omega$
$V_f = 1\,050$mV	$f_{ro} = 6{\cdot}6$Gc/s

The design is carried out in the following steps:

(1) *Value of R*

Using equation (4)

$$R_s + R < |R_j|$$
$$R < 2{\cdot}4$$

Let $R = 2{\cdot}0\Omega$

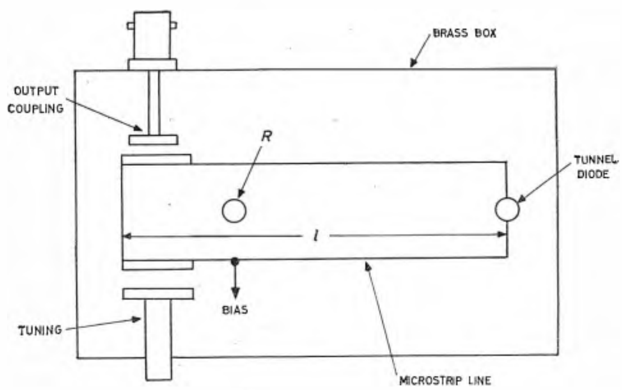
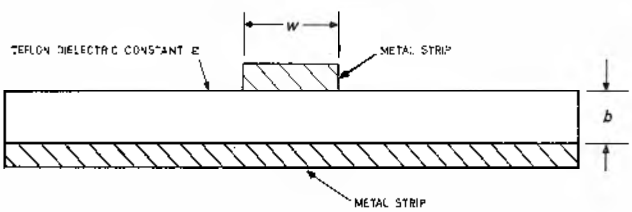


Fig. 7. A simple microwave oscillator

Fig. 8. Microstrip transmission line (cross-section)



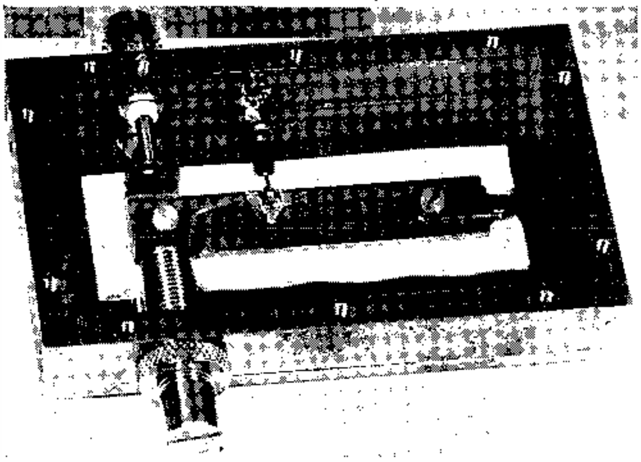


Fig. 9. A 2Gc/s tunnel diode oscillator

(2) *Value of* Z_0 (the characteristic impedance of the line.) Because of the small physical dimensions of the tunnel diode a microstrip resonator is used.

Referring to Fig. 8 Harvey[7] gives:

$$Z_0 = \frac{10^4}{3\sqrt{\epsilon}\,[7 + 8{\cdot}83\,.\,(w/b)]} \quad \ldots\ldots\ldots\ldots (7)$$

If $w = \frac{1}{2}$in, $b = 0{\cdot}010$in, $\sqrt{\epsilon} = \sqrt{2{\cdot}1}$, then $Z_0 = 5{\cdot}1\Omega$.

(3) *Wavelength in the line* (λ_g)

In microstrip the mode of propagation is not purely a TEM wave so that the wavelength in the line (λ_g) is not necessarily related to the free space wavelength (λ_0) by:

$$\lambda_g = \lambda_0/\sqrt{\epsilon} \quad \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

However, in the present case $b/w \rightarrow 0$ so that equation (8) is almost correct[7].

$$f = 2\text{Gc/s},\ \lambda_0 = 15{\cdot}0\text{cm}$$

$$\therefore\ \lambda_g = 10{\cdot}35\text{cm}.$$

(4) *Length of line* (l)

From equation (1) the admittance of the diode (Y_d) at 2Gc/s is found to be:

$$Y_d = -0{\cdot}11 - 0{\cdot}21\,\text{j} \quad \ldots\ldots\ldots\ldots\ldots (9)$$

Thus the resonator must present a capacitance susceptance of $+0{\cdot}21$ j. An open-circuited line somewhat longer than $\lambda_g/2$ will provide the correct susceptance and also provide a voltage null point for R.

From the Smith chart, $l = 6{\cdot}52$cm and the resistor R is placed 2·6cm from the open end (i.e. $\lambda_g/4$).

(5) *Power*

The theoretical power output is given by equation (6)

$$P \simeq 3/16\ (V_v - V_p)\ (I_p - I_v)$$

Using the values given in Table 1, $P \simeq 3{\cdot}4$mW.

(6) *Construction*

The oscillator may be built inside a metal box to shield the unit and provide mechanical rigidity for the output coupling and the tuning. The tunnel diode can be clamped to the microstrip with nylon screws thus avoiding the difficult and risky procedure of soldering such a small device.

A 2Gc/s tunnel diode oscillator is shown in Fig. 9, the design following the example above. The power output was 2·5mW and the stabilizing resistor used was of the ordinary $\frac{1}{2}$W carbon variety. Electronic tuning over a considerable frequency range is possible by placing a variable-capacitance diode across the end of the resonator.

Conclusion

The theory and design parameters of the tunnel diode have been examined in connexion with its application in microwave oscillators. The article includes full design information for a 2Gc/s oscillator and an extensive bibliography.

Acknowledgment

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REFERENCES

1. Esaki, L. A New Phenomenon in Narrow Germanium p-n Junctions. *Phys. Rev. 109*, 603 (1958).
2. Chow, W. F. Principles of Tunnel Diode Circuits. (Wiley, New York, 1964).
3. Gentile, S. P. Basic Theory and Application of Tunnel Diodes. (Van Nostrand, New Jersey, 1962).
4. Sterzer, F., Nelson, D. E. Tunnel diode Microwave Oscillators. *Proc. Inst. Radio Engrs. 49*, 744 (1961).
5. Microwave Solid-state Engineering. (Van Nostrand, New Jersey, 1964).
6. Hines, M. E. High-frequency Negative-resistance Circuit Principles of Esaki Diode Applications. *Bell Syst. Tech. J. 39*, 477 (1960).
7. Harvey, A. F. Microwave Engineering. (Academic Press, London, 1963).

BIBLIOGRAPHY

1. Sommers Jnr., H. S. Tunnel Diodes as High-frequency Devices. *Proc. Inst. Radio Engrs. 47*, 1201 (1959).
2. Rutz, R. F. A 3000Mc/s Lumped-parameter Oscillator Using an Esaki Negative-resistance Diode. *I.B.M. J. Res. Developm. 3*, 372 (1959).
3. Pulfer, J. K. Voltage Tuning in Tunnel-diode Oscillators. *Proc. Inst. Radio Engrs. 48*, 1155 (1960).
4. Trambarulo, R., Burrus, C. A. Esaki Diode Oscillators from 3 to 40kMc/s. *Proc. Inst. Radio Engrs. 48*, 1776 (1960).
5. Burrus, C. A. Millimeter Wave Esaki Diode Oscillators. *Proc. Inst. Radio Engrs. 48*, 2024 (1960).
6. Shelton, W. L., Frederick, W. L. Octave Tunable Tunnel Diode Oscillators. *Microwave J.*, 192 (Sept. 1962).
7. Hauer, W. B. A 4mW, 6kMc/s Tunnel Diode Oscillator. Digest of Technical Papers, 1962 Int. Solid-State Circuits Conf., p. 68.
8. Trambarulo, R. Some X-band Microwave Esaki-diode Circuits. Digest of Technical Papers, 1961 Int. Solid-State Circuits Conf., p. 18.
9. Sterzer, F. Stability of Tunnel-diode Oscillators. *RCA Rev. 23*, 396 (1962).
10. Kim, C. S., Brandli, A. High-frequency, High-power Operation of Tunnel Diodes. *I.R.E. Trans. CT-8*, 416 (1961).
11. Kim, C. S., Hopkins, J. B. High-frequency and High-power Operation of Tunnel Diodes. Digest of Technical Papers, 1961 Int. Solid-State Circuits Conf., p. 22.
12. Tarnary, K. Voltage Tuning of Tunnel Diode Oscillators. *Proc. Inst. Elect. Electronic Engrs. 51*, 385 (1963).
13. Davidsohn, U. S., Hwang, Y. C., Ober, G. B. Designing with Tunnel Diodes. *Electronic Design. 8* (Feb. 1960).
14. Smullen, L. I., Youla, D. C. Stability Criteria for Tunnel Diodes. *Proc. Inst. Radio Engrs. 49*, 1206 (1961).
15. Esaki, L., Miyahara, Y. A New Device Using the Tunnelling Process in Narrow p-n Junctions. *Solid State Electronics. 1*, 13 (March 1960).
16. Lesk, I. A., Holonyak, N., Davidsohn, U. S. The Tunnel Diode Circuits and Applications. *Electronics. 32*, 60 (Nov. 1959).
17. Sklar, B. The Tunnel Diode—Its Action and Properties. *Electronics. 32*, 54 (Nov. 1959).
18. Lesk, I. A., Holonyak, N., Davidsohn, U. S., Aarons, M. W. Germanium and Silicon Tunnel Diodes—Design, Operation and Application. *Wescon Conv. Rec. I.R.E.*, pt. 3, p. 19 (1959).
19. Price, P. J., Radcliffe, J. M. Esaki Tunnelling. *I.B.M. J. Res. Developm. 3*, 364 (1959).
20. Hall, R. N. Tunnel Diodes. *Trans. I.R.E. ED-7*, 1 (1960).
21. Holonyak, N., Lesk, I. A. Gallium Arsenide Tunnel Diodes. *Proc. Inst. Radio Engrs. 48*, 1405 (1960).
22. Murray, R. P. Biasing Methods for Tunnel Diodes. *Electronics. 33*, 82 (June 1960).
23. Gaertner, W. W. Esaki or Tunnel Diodes. *Semiconductor Products. 3*, Nos. 5. 6 (May, June 1960).
24. Hambleton, K. G., Low, J. J., Sherwell, R. J. Gallium Arsenide Tunnel Diodes. *Nature. 185*, 676 (1960).
25. Cox, L. G. Low Inductance Mount keeps Tunnel Diode From Oscillating. *Can. Electronics Engng. 4*, 31 (Sept. 1960).
26. Dill, H. G., MacPherson, M. R. Tracing Tunnel Diode Curves. *Electronics. 33*, 62 (Aug. 1960).
27. Lesk, I. A., Suran, J. J. Tunnel Diode Operation and Application. *Elect. Engng. 79*, 270 (1960).
28. Tenzer, M. Trends in Semiconductor Devices for Microwave Applications. *Electro-Technology. 68*, 82 (Oct. 1961).
29. Dacey, G. C. Esaki Diodes. *Bell Lab. Rec. 39*, 122 (April 1961).
30. Burrus, C. A. Gallium Arsenide Esaki Diodes for High-frequency Applications. *J. Appl. Phys. 32*, 1031 (1961).
31. Thompson, P. M., Bateson, J. Introduction to Tunnel Diode as Circuit Element. *J. Brit. Instn. Radio Engrs. 22*, 75 (1961).
32. Hauer, W. B. Definition and Determination of Series Inductance of Tunnel Diodes. *Trans. I.R.E. ED-8*, 470 (1961).
33. Przybylski, J., Roberts, G. N. Design and Construction of Tunnel Diodes. *J. Brit. Instn. Radio Engrs. 22*, 497 (1961).
34. Ishii, K., Hoffins, C. C. Extending Tunnel Diode Operating Frequency. *Electronics. 35*, 43 (June 1962).
35. Jonscher, A. K. Physics of Tunnel Diode. *J. Appl. Phys. 12*, 654 (1961).
36. Cristal, E. G. Proposed Technique for Stabilization of Tunnel Diodes. *Microwave J. 5*, 108 (1962).

Multi-Aperture Devices for All-Magnetic Logic Applications

By D. J. Morris*, Dipl. Eng., M.Sc., A.M.I.E.E., A.M.I.E.R.E.

This article describes the performance of a number of multi-aperture devices of different shapes designed for the basic combinational logic operations. The devices described here may be combined to perform sequential logical operations with no additional components required beyond the use of wire for connecting the devices. The system used for coupling the devices is known as m.a.d.-R. It allows wide operational tolerances and is intended for application where high reliability is required.

(Voir page 564 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 571)

IN the past few years research has been carried out to design all-magnetic logic systems which incorporate multi-aperture devices and the connecting wire only. Transistors and other associated components are not used within the actual logic circuits, although they are still required for the pulse sources which are essential for driving these devices. The use of only magnetic devices in logic circuits is highly desirable owing to the high reliability and the long life performance of these components.

Multi-aperture devices are made of the same material as the ferrite cores used in computer stores, and as such possess the same rectangular hysteresis properties. In ferrite cores switching is controlled by the threshold m.m.f., while in multi-aperture devices the switching is controlled by the minimum cross-sectional area of the switching paths. In a selected path, no remanent flux switching can take place by induced fields smaller than the coercive force but its flux may switch with the coincidence of two induced fields, both of magnitude slightly less than the coercive force. This characteristic is used in the selection of ferrite cores in a computer store. On the other hand, there could be no further switching once the flux in the minimum cross-sectional area has been switched; increasing the induced field above this limit will not change the amount of flux switching but will reduce the time required for the flux to switch. Thus, in multi-aperture devices, the only limit to the driving fields is the lower one, which must be of sufficient magnitude to switch the flux in the minimum cross-sectional area of the selected path, and hence this gives wide operational tolerances. In multi-aperture devices there may be a large number of switching paths, which enable various logical operations to be performed by a single device, while in the ferrite cores there is only one switching path. Furthermore, the different switching paths within these devices can provide the isolation between input and output circuits; hence, by eliminating the need for buffers, they are ideally suited for all-magnetic logic operations. Consequently, the devices can be heavily loaded by very low impedance coupling loops without reflecting the output load to the input circuit.

This article describes the respective shapes of a number of multi-aperture devices which can perform various combinational logical operations. The devices which were selected exhibit the feature of having the output circuit isolated from the input circuit, thus enabling them to be assembled together to perform sequential logic operations with only the connecting wires. Devices, such as the Laddic, which may be classed as a form of multi-aperture device are not discussed here since they do not exhibit the required isolation feature.

* *English Electric-Leo-Marconi Computers Ltd.*

Although some of the shapes described here have been presented elsewhere, very little has been published about their performance. Further study in this field is therefore called for. The purpose of this article is to collect a set of major devices, modifying their configuration and explain their performance and also to introduce new shapes. Most of these devices were not available for investigation; nevertheless, flux patterns and mode of operation, based on the observed behaviour of other shapes, are suggested.

The multi-aperture devices selected here are intended for 'fail-safe' operations. These are operations in which

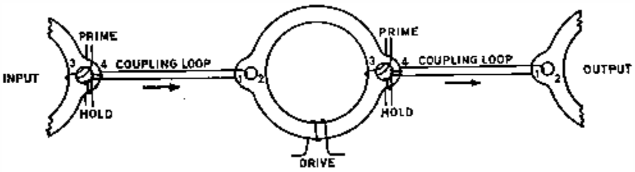


Fig. 1. The basic transfer operation

failure of any of the applied m.m.f.'s or any other parameter will cancel the output pulse. This accounts for the exclusion of those shapes which require holding currents which are applied to specific apertures in order to perform the required logical operation. (Failure of this 'holding' current may produce an output pulse for an unwanted input condition).

There are a number of alternative all-magnetic systems. The system adopted here is one known as m.a.d.-R[1] chosen for its greater reliability and fail-safe considerations. The distinctive name of this system derives from the initial of the words multi-aperture device resistance. In this system the resistance of the wire coupling the multi-aperture devices is of extreme importance[2]. The operational tolerances of the m.a.d.-R system are comparatively wide in relation to other all-magnetic systems, although the speed of operation is relatively slow.

The Basic Operation

The m.a.d.-R all-magnetic logic system is best explained by describing the basic transfer circuit, shown in Fig. 1, used for shift registers. A three-aperture device is used having two small apertures for the input and output and a large aperture for the control of the whole device. This forms, on the X axis, four distinct legs; legs 1 and 2 around the input aperture and legs 3 and 4 around the output aperture. Before one can discuss this circuit and its performance, the different flux states of the device must be considered, and these are shown in Fig. 2.

In the initial state of the device, known as the 'block' state and shown in Fig. 2(a), the flux round the large aperture is in a clockwise direction, i.e., the flux is in an

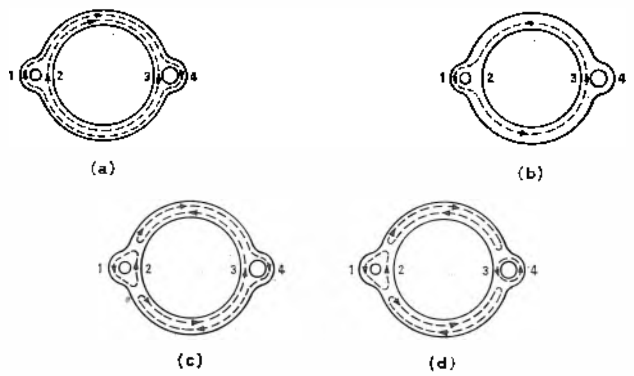


Fig. 2. Flux patterns of the transfer device
(*a*) *Block state*
(*b*) *Flux switching path during input-set*
(*c*) *Set state*
(*d*) *Prime state*

upward direction in legs 1 and 2 and in a downward direction in legs 3 and 4. This state of the device is formed by inducing a field in the drive winding round the large control aperture.

An input field, induced in a winding round leg 1, will tend to switch the flux in that leg in a downward direction. Flux in square loop materials always switches in closed loops and along the shortest available path. The shortest return path for a field induced in leg 1 is along leg 2. However, the flux in leg 2 is not 'available', since the fluxes in legs 1 and 2 (when the device is in its initial state) are in opposite directions relative to a field induced in the input aperture. Hence, the input m.m.f. will cause the flux in leg 2 to be driven further into the saturation but will not switch it. If the magnitude of the induced input field is large enough, the flux will switch round another path which is round the large control aperture.

This is possible since the flux is in the same direction relative to a path round this large aperture. The flux can switch along the return path provided by leg 3 and leg 4, but as leg 3 provides the shortest available path it will switch along this leg. Hence, an input field induced in leg 1 will reverse the flux in a counter clockwise direction round the large aperture, as shown in Fig. 2(b), i.e. in a downward direction in leg 1 and in an upward direction in leg 3. After the termination of the input field, there will be some further re-orientation of the domains round the input aperture, so as to preserve flux continuity, as shown in Fig. 2(c). The device in this state is known as the 'set' state.

Once the device is set, a field induced in the output aperture, known as the prime field, may switch the flux round this aperture and thus reverse the flux in an upward direction in leg 4 and in a downward direction in leg 3, as shown in Fig. 2(d). The magnitude of the induced prime field must be limited so that it switches the flux only round the output aperture (along legs 3 and 4) but not round the large aperture. The prime field may be induced by a pulse or a constant current. A constant current is preferable because it eliminates the need for extra clock pulses. In the block state, the prime field cannot switch any flux round the output aperture, as the fluxes in legs 3 and 4 are in opposite directions relative to the field round this aperture. There is a danger, however, that a large magnitude prime field may switch the flux round the large aperture thus setting the device while it is in its block state; hence the prime magnitude must be limited. The prime operation (having the prime field within its limits) may take effect only after the device is set, and this is the reason for calling the two states of the device 'block' and 'set', terms which may also be used to represent the binary information of '0' and '1'.

When the device is switched back to its blocked state, after first being switched to its set and prime states, an output pulse will be induced in a winding round leg 4 as the flux switches from its upward direction back to its downward direction.

The Transfer Operation

The various switching paths within the multi-aperture device allow at least two magnetic switching circuits within each device. These two internal magnetic circuits can be used to provide input-output isolation, so that the device can be heavily loaded by very low impedance coupling loops. In the device described in the previous section the input is round the large aperture along legs 1 and 3, and the output circuit is along legs 3 and 4. Hence, a load on leg 4 is isolated from leg 1 and the output load is not reflected to the input circuit. The prime operation connects both circuits, by transferring the information from the input magnetic circuit to the output circuit.

Returning to the basic circuit shown in Fig. 1, a current in the input coupling loop will cause the device to switch from its blocked state to its set state. The low impedance of the output coupling loop is not reflected during the input operation, but, during the prime operation, the output load will heavily damp the device.

During the prime operation there will be flux switching in leg 4, giving an output voltage pulse which will consequently cause a current to flow in the output coupling loop. This current will not affect the following device since it flows in a direction which will only tend to drive leg 1, of the following device, further into saturation Therefore the prime operation takes the following form:

$$N_T \frac{d\Phi_4}{dt} + IR = 0$$

having the resistance of the coupling loop wire as R the current in the loops as I and the number of turns N_T on the output winding.

The resistance of the wire must be greater than zero so as to enable the flux to switch leg 4. Unfortunately the resistance must still be small, of the order of a fraction of an ohm, so as to enable the flux transfer from one device to another; also, the number of turns must be small. As explained in the previous section, the magnitude of the prime field must be limited and thus current will be small. The only remaining variable factor in the above function is the switching time, which will be of a relatively long duration because of the low resistance. To increase the magnitude of the permitted prime field and hence speed this operation, the prime winding is shared between legs 3 and 4.

After the prime operation a pulse in the drive winding round the large aperture will drive the device back to its block state.

This operation will also produce flux switching in leg 4 and give an output voltage pulse having a polarity opposite to that obtained from the prime operation. This output pulse will cause a current to flow in the coupling loop, this time, in a direction which switches the flux in leg 1 of the following device. Thus the information is transferred from one device to the following device. During this operation there will be flux switching in both devices, which is governed by the following:

$$N_T \frac{d\Phi_4}{dt} + IR + N_R \frac{d\Phi_1}{dt} = 0$$

where N_R is the number of turns on the input winding.

It can be seen that best transfer is achieved when the ·esistance is equal to zero; i.e. when

$$|\Delta \Phi_4| \rightarrow \left| \frac{N_R}{N_T} \Delta \Phi_1 \right|$$

The resistance value of the coupling loop wire is a :ompromise between the requirements of the prime and .he drive operations; i.e. the resistance must be extremely ;mall but larger than zero.

During the drive operation, the device switches from ts prime state, Fig. 2(d), to its block state, Fig. 2(a). It :an be seen that there will be flux switching not only n leg 4 but also in leg 1. This will cause a current to low in the input coupling loop from the device in |uestion to the previous device. In order to prevent his back flow of information, the flux in leg 4 of the ırevious device must be held while the device in question ; driven to its blocked state.

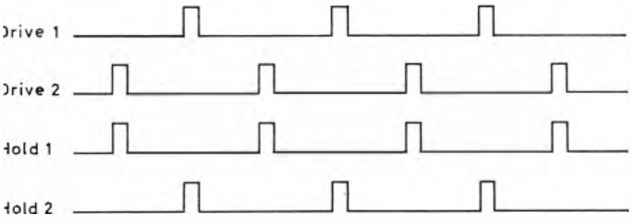


Fig. 3. Schematic arrangement of the clock pulses

The information is processed and transmitted from de- ice to device in a step by step propagation. Since a evice cannot both receive and transmit information at ıe same time, two clock drive pulses must be used, as ıown in Fig. 3. One clock pulse corresponds with the me the device is set, while the second clock pulse is sed for driving it to its blocked state. The drive pulse pplied to one device must also be applied, at the same me, to the hold winding on leg 4 of the previous device.

he Basic Shape of the Device

The shape of the device plays an important part in etermining the range of operation of the device. The ide tolerance of the driving m.m.f.'s are due to the ıriations in the geometry of the device.

The ratio of the diameters of the large and small aper- :res must be large, so as to reduce the danger of the :ime field setting the device during its blocked state. he larger the ratio, the larger will be the permitted prime .m.f., and hence the faster the speeds which are ob- ined. On the other hand, as this diameter ratio is in- eased, the flux path length round the larger aperture also increased and more m.m.f. would be required r setting and blocking the device.

The switching of flux is governed by the minimum oss-sectional area of the switching path. Outside the nge of the switching paths there may be regions of cal flux closures where the direction of flux does not llow the required pattern. In order to prevent these cal flux regions and to have equal cross-sectional areas und the large aperture, the material round the minor ıertures is made in an ear shape form. Furthermore, eliminate any of these regions round the output aper- re, the cross-sectional area of the legs round that ıerture must be smaller than the cross-sectional area of put aperture legs.

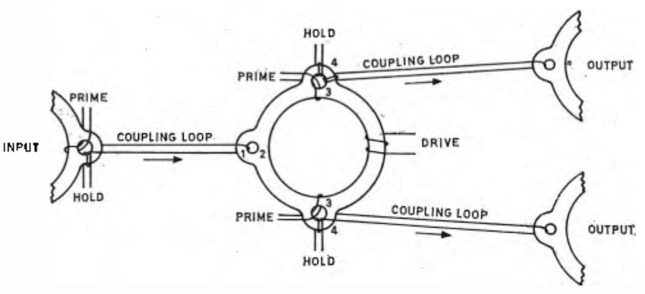


Fig. 4. The fan-out operation

The reasons which determine the specific configuration of the basic multi-aperture device were discussed in a previous publication[3]. These restrictions, especially those employed on the input and output apertures, apply to other shapes too, and have been used for the design of all the other shapes discussed in this article.

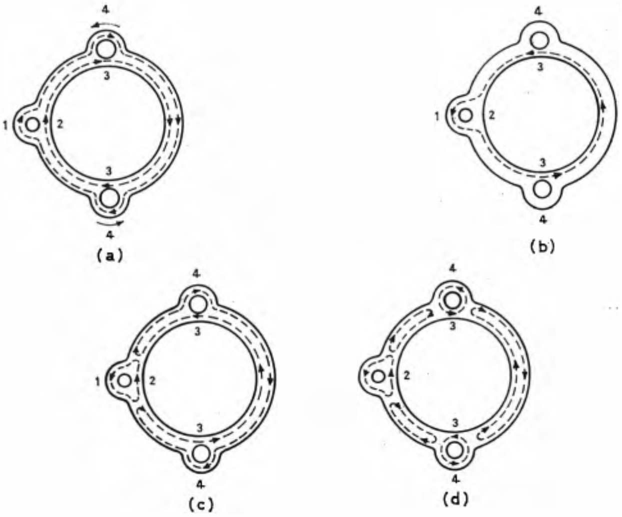
The Fan-Out Operation

In the previous sections, the devices were described as having a single input and a single output. In order to utilize effectively the information which is processed in the device, methods must be found to feed-out this in- formation to a number of other devices. This operation, known as the fan-out operation, may be achieved by having one device with a single input and a number of outputs.

The fan-out operation may be obtained by having a number of coupling loops assembled on the output aper- ture. However, the power available from a single output aperture is limited and is only sufficient to transfer flux to one other device. Furthermore, as the number of coupling loops is increased, the reflected load during the prime operation will consequently be reduced, bringing it to such a low impedance so as to prevent the prime operation from functioning. Hence, a condition must be imposed so that the output aperture of one device may be linked to only one other device, with one coupling loop which has a perfect loop resistance.

In order to achieve the fan-out operation, a new device

Fig. 5. Flux patterns of the fan-out device
(a) *Block state*
(b) *Flux switching path during input-set*
(c) *Set state*
(d) *Prime state*



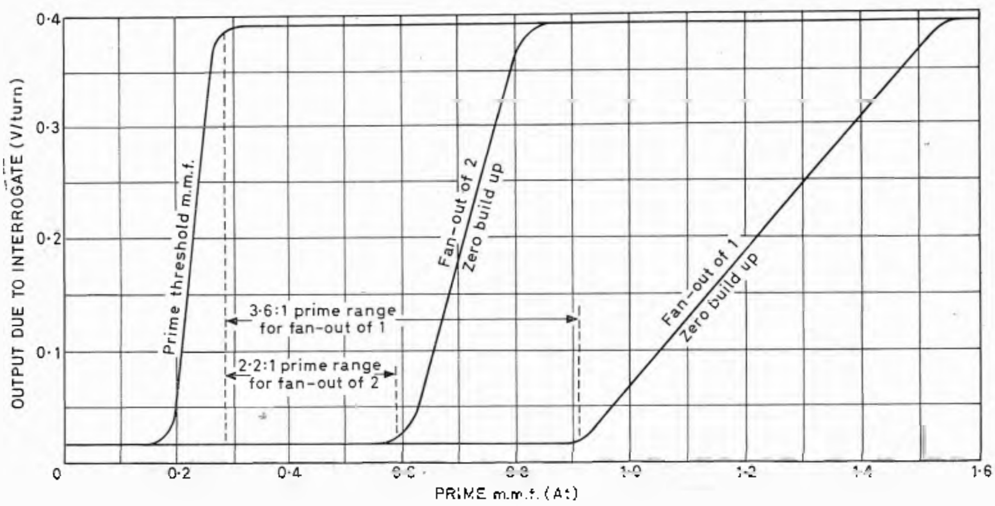


Fig. 6. Range of prime m.m.f. in fan-out device

must be used which has one input aperture and a number of output apertures. Such a device, which has the capabilities of a fan-out of two; is shown in Fig. 4. Each of the two output apertures is primed separately, and held separately to prevent back flow of information. The two reflected loads are each isolated from the input aperture and also from each other.

The flux patterns of this device, the performance of which is similar to that of the transfer device, are shown in Fig. 5. The operation of each of the two output apertures is exactly the same as that for the single output aperture, as explained before. This is why the legs were numbered 3 and 4 in both output apertures.

The set m.m.f. switches the flux in both of the legs marked 3—as shown in Fig. 5(b)—but will be of the same order of magnitude as was applied to the transfer device. This is due to the fact that the path length is the same for both devices and no load is reflected during the input operation. The prime operation will occur locally round each aperture and the duration of this operation will be the same as for the transfer device, since the operation occurs simultaneously in both output apertures. The only basic difference between the fan-out and transfer devices is that extra drive m.m.f. must be supplied to the fan-out device so as to enable it to cope with the extra load when it feeds out its information to the other devices.

Priming both output apertures (by d.c. m.m.f.'s) causes new practical difficulties, as the operational range of the prime m.m.f. is reduced. There is a danger that the two prime fields induced in both legs 4, may add up and set the device during its blocked state. This danger is indicated by the direction of the arrows at the side of the device shown in Fig. 5(a). Similarly, the fields induced in both legs 3 may also add up and block the device during its set state. The curves shown in Fig. 6 demonstrate the change in the range of the prime m.m.f., when priming one and two output apertures.

These curves were obtained in an experiment using two devices (connected by a wire-only coupling loop), varying the prime applied to the first device and recording the output (due to the interrogation) of the second device. The lower limit indicates the region where no information is transferred to the second device, in other words, a binary one is transferred as a zero. The upper range indicates where a zero in the first device is transmitted as a one.

Since the range of the prime m.m.f. is narrowed, with the increase of the number of primed output apertures, it is not practical to have a single device giving a fan-out

Fig. 7. The OR operation

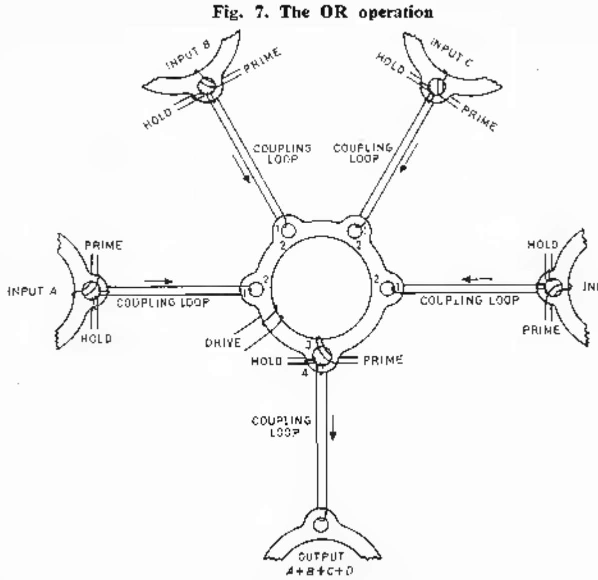
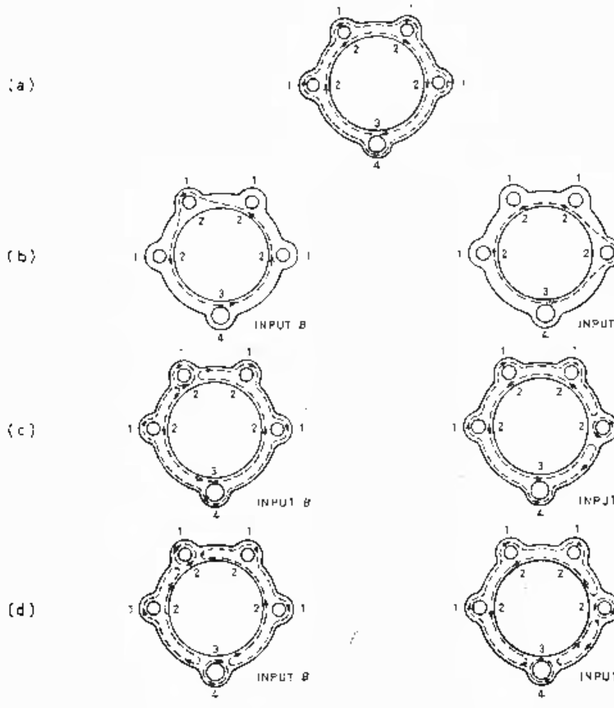


Fig. 8. Flux patterns of the OR device

(*a*) *Block state*
(*b*) *Flux switching paths during input-set*
(*c*) *Set state*
(*d*) *Prime state*



of more than two or at the most three. The range of 2:1 obtained for a fan-out of two may still be reduced when consideration is given to temperature changes and material tolerances.

The 'OR' Operation

As in the case of the fan-out operation, it is impossible to feed-in the information from a number of coupling loops assembled together on one input aperture to achieve

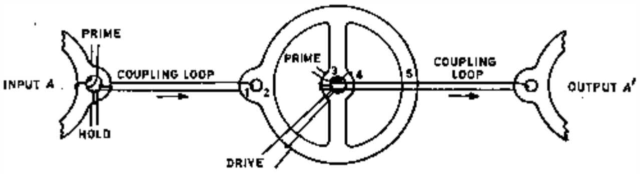


Fig. 9. The complement operation

the OR function. Only one coupling loop may be connected to each input aperture since a second low impedance coupling loop will damp the input leg 1 and prevent it from switching. It was shown before that the output load is isolated from the input circuit when the device is set. Deprived of this isolation feature, not only the device in question will be prevented from switching but also the input device.

It was shown that the transfer of information from one input device to the next device takes the following form:

$$N_T \frac{d\Phi_4}{dt} + IR + N_R \frac{d\Phi_1}{dt} = 0$$

If a secondary winding is assembled on leg 1 of the input aperture, an output voltage pulse will be induced across it as the flux in leg 1 changes when the device is set. If a second coupling loop (from a second input device) is connected to the same leg 1, there will be a current flowing in this loop but no flux switching in this second

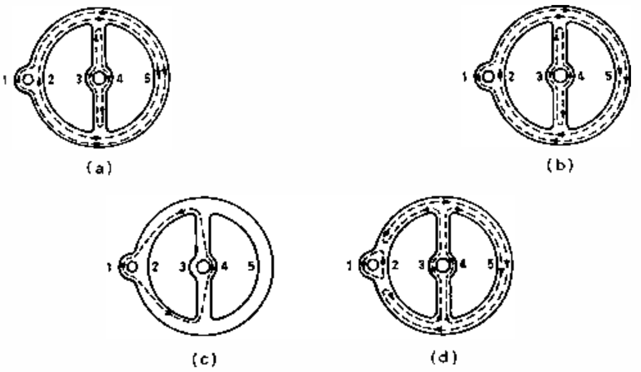


Fig. 10. Flux patterns of the complement device

(*a*) *Reset state*
(*b*) *Prime state (with no input)*
(*c*) *Flux switching path during input*
(*d*) *Set state*

input device. Hence, the function in this second coupling loop will take the following form:

$$IR + N_R \frac{d\Phi_1}{dt} = 0$$

Since the second function is caused by the change of the flux in leg 4 of the first input device, it is obvious that $\Delta \Phi_4 = 0$.

To achieve the OR function, a special input aperture is required for each input coupling loop, and a magnetic circuit which performs this operation is shown in Fig. 7.

The number of input apertures is not limited, as in the fan-out device, since the input apertures are not primed and each input aperture is isolated from another, as may be seen from the flux patterns given in Fig. 8. The setting flux switching paths will be along the shortest available paths, which will always be along the required leg 3 and along the other legs 2. This is demonstrated by two examples, one having input B and the other input D.

If inputs are applied simultaneously to two (or more) input apertures, then the flux will switch along both respective legs 1 and along leg 3. Hence, this device performs the INCLUSIVE-OR function.

The devices used for the fan-out and for the OR operations both have the cross-sectional area of the material round the output apertures smaller than the input apertures.

The Complement Operation

In the previous sections the devices produced an output as a direct function of the input. An important logical

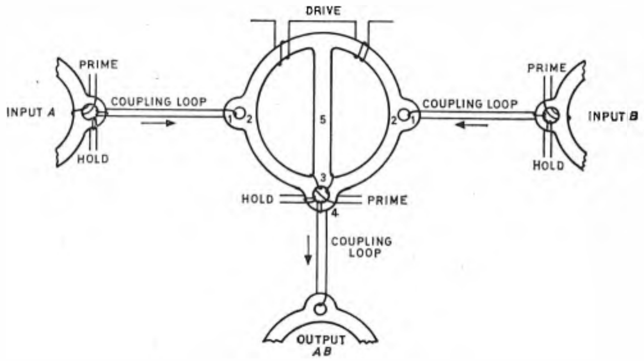


Fig. 11. The AND operation

function is the complement of this operation, which only gives an output when no input is applied. This operation is also known in the literature as negation, NOT and negative transfers. If the previous operations produce a positive transfer the complement transfer may be called a negative transfer.

The circuit which performs the complement operation is shown in Fig. 9. A new shaped device is used which has a central limb with an output aperture in it. The input is similar to the positive transfer but the output is taken from leg 3 of the centre aperture. The hold winding, preventing back flow of information, is also situated on leg 3, but due to shortage of space is not shown in Fig. 9.

The flux patterns of the complement device are shown in Fig. 10. The drive m.m.f. is applied in the centre output aperture, which switches the flux round the whole device in a clockwise direction, as shown in Fig. 10(a). The prime m.m.f. may switch the flux round the centre aperture in this initial state of the device with no input applied, and for this reason this state is called 'reset' and not 'blocked'. When the next drive pulse is applied, after the device has been switched to its prime state, as shown in Fig. 10(b), an output pulse will be induced in the winding assembled on leg 3.

If an input is applied to leg 1 when the device is in its reset state the flux will switch along legs 1 and 4, as shown in Fig. 10(c), since leg 4 provides the shortest available path. In this so-called 'set' state, shown in Fig. 10(d), the prime m.m.f. cannot switch any flux round the output aperture, as the flux in this state is saturated in opposite directions relative to the prime field. In effect the complement device in this state is blocked.

The output from this device is obtained only when no input is applied, thus representing the Boolean Algebra notable $\bar{A}$. This device was not available for investigations, but it is believed that the prime m.m.f. would be required to be in pulse form and not d.c. as used before. This is to ensure that when the input m.m.f. is applied the flux pattern in the device will be as shown in Fig. 10(a) and not as in Fig. 10(b).

The 'AND' Operation

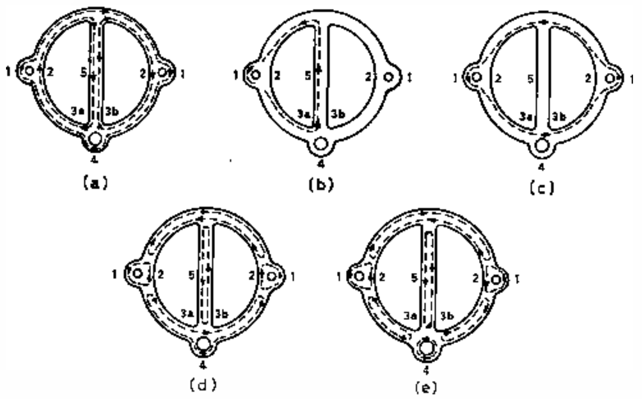
The AND operation may be performed by a single device as shown in Fig. 11. This device is similar to the complement device by having a centre limb and two large apertures. However, it has two input apertures situated at each side of the device and an output aperture at one side of the centre limb.

The flux patterns of the AND device are shown in Fig. 12. The drive windings are round the two large apertures and thus the drive m.m.f. will switch the flux in a clockwise direction round each of the large apertures. In this state of the device, shown in Fig. 12(a), the device is blocked, since the prime m.m.f. cannot switch any flux round the output aperture.

If only one input m.m.f. is applied to the device, it will switch along the centre limb as shown in Fig. 12(b). In this state too, the prime m.m.f. will be unable to switch the flux round the output aperture, since part of leg 3 is still in the opposite direction to the flux in leg 4 (relative to a path round that aperture). Only after both input m.m.f.'s are applied, as shown in Fig. 12(d), will the flux in leg 4 be in the same direction as that in leg 3, thus enabling the prime operation to take place and bringing the device to the state shown in Fig. 12(e).

The flux switching path, when both inputs are applied, shown in Fig. 12(c), is of an interesting nature. If only one input is applied, either only A or only B, the return path is always along the centre limb, since this provides the shortest available path. When both inputs are applied, there will be no switching at all in the centre limb. The reason for this is that when both inputs are applied simultaneously, the two induced fields will combine together and switch along the outer rim of the device, which is shorter than the sum of the two paths round each of the two large apertures.

Fig. 12. Flux patterns of the AND device
(a) Block state
(b) Flux switching path during input AB
(c) Flux switching path during input AB
(d) Set state (after input $A\bar{B}$)
(e) Prime state



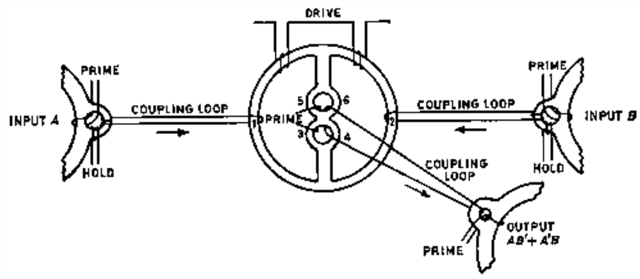


Fig. 13. The exclusive-OR operation

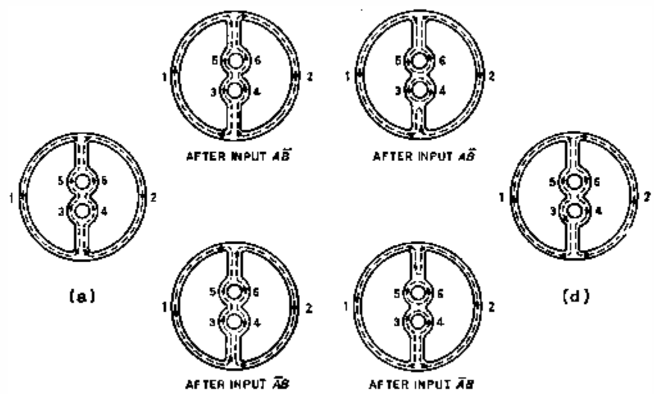


Fig. 14. Flux patterns of the exclusive-OR device
(a) Block state
(b) Set state
(c) Prime state
(d) Switched state

The prime winding must be shared between leg 4 and both halves of leg 3 so as to eliminate any undesired switching when only one input is applied

Exclusive-OR Operation

The EXCLUSIVE-OR device, shown in Fig. 13, differs from all the devices discussed previously, in having large input apertures. There are also two output apertures and the output is taken from a winding round both apertures.

The flux patterns of this device are shown in Fig. 14. In the initial state of the device, shown in Fig. 14(a), the flux in the centre limb is in an upward direction and in the two outer limbs of the device in a downward direction. The prime, which induces a field round both output apertures, cannot switch any flux in either aperture, since the flux is in opposite directions round each of the two apertures. If the input is applied to either of the two inputs, the flux will switch in half the centre limb, as shown in Fig. 14(b). An input to limb 1 will switch the flux along legs 3 and 5, and an input to limb 2 will switch along legs 4 and 6. In either of these two set states, the prime field can switch the flux round one of the output apertures. For an input to limb 1, the prime will switch in clockwise direction along legs 3 and 4, and for an input to limb 2, the prime will switch in a counter clockwise direction along legs 5 and 6, as shown in Fig. 14(c). The direction of the output pulses obtained from the device, when it switches from its prime state back to its blocked state, will be of the same polarity whether the previous input was on limb 1 or on limb 2.

In this device it is possible to reverse the direction of the prime m.m.f. without affecting the operation of the device, although the output pulses are with an opposite polarity to that obtained before. In all the other devices discussed here, the direction of the prime m.m.f. must be in one specific direction. If the direction of the prime applied to this device is reversed, it will switch in a clockwise direction round legs 5 and 6 for an input to limb 1, and in a counter clockwise direction round legs 3 and 4 for input to limb 2.

If both inputs are applied to the device, the flux in both halves of the centre limb will switch in a downward direction, as shown in Fig. 14(d). In this state of the device, the prime m.m.f. cannot cause any flux switching round any of the two output apertures.

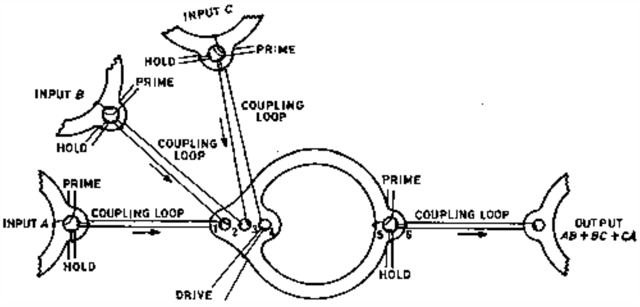


Fig. 15. The majority-decision operation

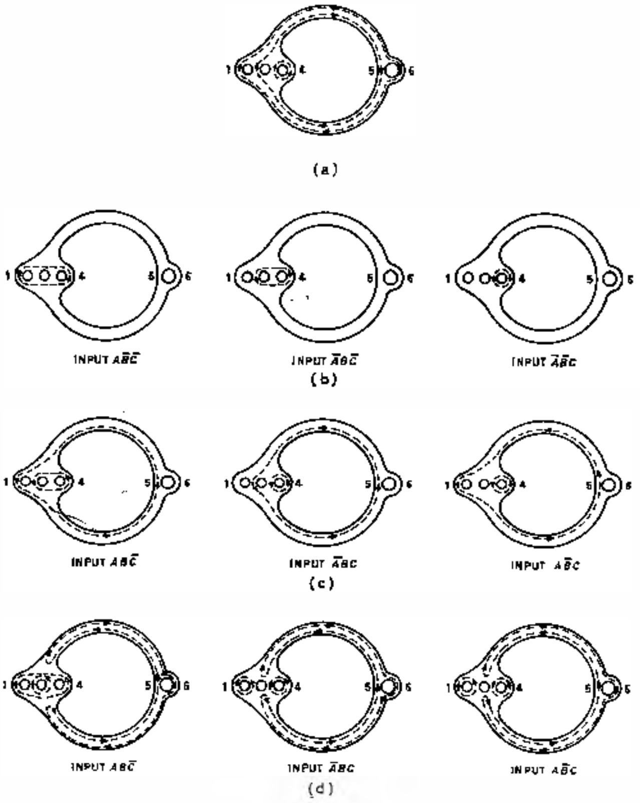


Fig. 16. Flux patterns of the majority-decision device

(*a*) *Block state*

(*b*) *Flux switching paths during a single input*

(*c*) *Flux switching during a double input*

(*d*) *Set state*

Previous experience with a similar device did not give favourable results, but this could have been due to the imperfect shape of that device. The shape shown in Figs. 13 and 14 has a modified configuration, which, it is hoped, improves on the performance of less efficient shapes. Since two output apertures are used, care must be taken to eliminate any local flux regions round these two apertures. This may be done by shaping the material round the output apertures, as shown in Figs. 13 and 14. Also the cross-sectional area of legs 3, 4, 5, and 6 must be equal to each other and the cross-sectional area between the two output apertures must be equal to twice the cross-sectional area of each leg. Likewise, the cross-sectional area of any of the inner legs. This last condition is to ensure that when both inputs are applied the flux in the centre limb will be fully saturated in a downward direction.

This configuration of the EXCLUSIVE-OR device suffers from the disadvantage that it is impossible to hold the output apertures so as to prevent back flow of information. A field induced in the output apertures, when the device is in its block state or in the state after input *AB*, will change the flux pattern in the centre limb rather than holding it in the required direction. In order to prevent back flow of information, the input aperture of the following device must be primed as shown in Fig. 13.

The Majority-Decision Operation

Although the majority-decision operation is not among the major logical operations, it was thought advisable to add it here so as to show some further complex logical operations which may be produced by a single device. This device, which is shown in Fig. 15, produces an output pulse if at least two out of its three inputs have been

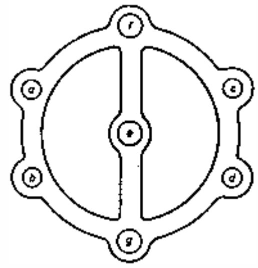


Fig. 17. A combined logic device

applied. It may also be used as a 'carry' device for a full-adder since it performs a similar logical function.

This device has three input apertures and one output aperture which forms six legs on the *X* axis. The three inputs are applied to legs 1, 2 and 3, while the output is taken from leg 6. The initial flux pattern of the device shown in Fig. 16(a), is achieved by inducing a field in the aperture between legs 3 and 4. This field will switch the flux round the whole device in a clockwise direction round the aperture between legs 3 and 4. Since the range of field is not limited by the cross-section of the legs it will form concentric flux lines round that aperture. This will cause the flux in legs 1, 2 and 3 to be in an upward direction and in legs 4, 5 and 6 in a downward direction.

A single input field induced in either of the three legs will switch the flux round its shortest available path, which will always be along leg 4, as shown in Fig. 16(b). No switching will occur in leg 5 and thus the prime operation cannot take place. The amount of material switched in this input operation is controlled by the minimum cross-sectional area of each of the input legs, irrespective of the magnitude of the induced input field.

If two input fields are induced, it will cause switching in both leg 4 and leg 5, as shown in Fig. 16(c). As explained before, there will be some further orientation round the input apertures, after the continuity; this, however, will not affect the direction of the flux in each of the six legs. The three different flux patterns, shown in

Fig. 16(d), represent the set state of the device since in all three cases the prime may switch the flux round the output aperture.

As in the previous devices, the cross-sectional area of the legs round the input apertures must be larger than the legs round the output aperture.

Conclusions

Although work on all-magnetic logic circuits has been going on for some years, both in the U.S.A. and here, there has been very little field experience in such circuits. Nevertheless, this sphere of operation shows promising results for applications where high reliability is required. However, as may be seen from this article, an excessively large number of different shaped devices is required to perform the logical operations. Research has yet to be carried out to design devices of simple shapes which may perform a wide variety of combinational logical functions. This may be done by combining a number of different shaped devices into one configuration. An example of such a device is shown in Fig. 17, which has four input apertures (*a*, *b*, *c* and *d*) and three output apertures (*e*, *f* and *g*). The device may be used as a complement device with two inputs applied to apertures *a* and *b* and the output obtained from aperture *e*, thus operating as a NOR device. If only one half of the device is used (one large aperture) it may be employed as a transfer or OR device. It may also be employed as an AND device which has a fan-out of two which is obtained from output apertures *f* and *g*. Another application of the device could be as OR-AND where the OR operation is carried out round each half of the device. This device is only an example of what can be done by combining different shapes. It is hoped that new devices will be produced in the near future which will avoid the necessity for a large number of different shapes.

Acknowledgments

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REFERENCES

1. BENNION, D. R. MAD—Resistance Type Magnetic Shift Register. Proceeding of the special technical conference on non-linear magnetics and magnetic amplifiers (A.I.E.E.), p. 96 (1960).
2. MORRIS, D. J. Performance of an All-magnetic Shift Register. *Proc. Instn. Elect. Engrs. 111*, 291 (1964).
3. MORRIS, D. J. The Shape and Performance of Multi-aperture Magnetic Device. *Electronic Engng. 37*, 440 (1964).
4. RAJCHMAN, J. A., LO, A. W. The Transfluxor. *Proc. Inst. Radio Engrs. 44*, 321 (1956).
5. CRANE, H. D. A High-Speed Logic System using Magnetic Elements and Connecting Wire Only. *Proc. Inst. Radio Engrs. 47*, 63 (1959).
6. COHLER, E. U., BAKER, T. E. Geometric Factors in Multi-aperture Ferrite Devices. Special technical conference on non-linear magnetics proceedings, Los Angeles, California (I.R.E. and A.I.E.E.), p. 215 (1961).
7. GIANOLA, U. F. All-magnetic Logic. *Solid State J. 2*, 21 (August 1961) (also in Bell Telephone System Monograph 3934).
8. BENNION, D. R. Research on Multi-aperture Magnetic Logic Devices. Stanford Research Institute, Project 2697, Technical Report 1 for Office of Naval Research, Contract 2712(00).

The Mullard Cryomagnetic Laboratory

The Mullard Cryomagnetic Laboratory at the Clarendon Laboratory, Oxford, was formally inaugurated on 30 June by Lord Bowden, Minister of State for Education and Science. The laboratory was made possible by a gift of £25 000 from Mullard Ltd. This event marks the completion of a major research facility whose origins go back to the 1940's when support was first given by the former Department of Scientific and Industrial Research. The new Laboratory will enable a number of research projects to be carried out simultaneously using the same basic equipment.

The name 'cryomagnetic' (cold and magnetic) is justified by three important connexions between low temperature and high magnetic fields: with strong magnetic fields low temperatures can be produced (magnetic cooling, nuclear cooling); with low temperatures strong magnetic fields can be produced more easily (e.g. super-conducting solenoids) and finally low temperatures and high magnetic fields go hand in hand to remove molecular disorder.

With the original DSIR grants the two rather distant scientific disciplines of nuclear physics and low temperature studies were brought together for the first time at the Clarendon Laboratory. Between 1948 and 1957 the Department granted about £38 000 for this particular work. In 1959 Dr. N. Kurti, FRS, a pioneer worker in this field, applied to DSIR for a grant to finance a new high field installation based on the existing 2MW motor generator to study nuclear orientation and nuclear cooling in magnetic fields up to 140kG.

This grant (which with later supplements came to nearly £40 000) provided for staff and equipment for a magnet laboratory having seven magnet stations.

The new installation replaces equipment commissioned in 1950 and is being used for a wide range of research topics in addition to nuclear orientation and nuclear cooling.

The Mullard Cryomagnetic Laboratory is built above the new Simon Workshop and occupies the whole of the first and second floors of the new building (approximately 200 square metres on each floor). A small plant room has been built on the third floor. On each of the first and second floors magnet stations are grouped round a central hall, in which there is a console carrying all the instruments and controls needed for operating the magnets. The console is sufficiently mobile to be moved into positions outside each room so that the operator can have a good view of the apparatus in the room, while at the controls.

On the first floor one of the four rooms grouped round the hall is used for the counting equipment required for the nuclear orientation experiments, and so there are only three magnet stations. On the second floor all four rooms round the hall contain high-power magnets.

Apart from the magnet laboratories, the building contains a number of offices for the permanent members of the staff, two offices for visiting research workers, and a workshop where specialized parts of the magnet are made, where maintenance is done, and where the magnets can be tested before being passed out for use in the research rooms. For the last purpose the full services of the 2MW generator and cooling system are provided in the workshop.

The 2MW motor-generator set was built in 1919 for the Corporation of Manchester, and until about 1935 it was used in connexion with their d.c. lighting system. When the city joined the national a.c. grid the machine became obsolete. It was bought by the Clarendon Laboratory in 1946 for £1 000 with the help of a grant from the Royal Society.

The set is driven by a 2 900 h.p. induction motor energized from the national grid at 6·6kV. Two 1 000kW generators are connected in parallel and are rated to deliver up to 4 500A at 460V. The set has been run at 10 per cent overload, and it is hoped that this can be increased, without trouble, to 25 per cent.

The cooling water is conveyed to the research rooms in 15cm diameter copper mains. The system has been designed to provide a maximum water flow of 2·5m^3/min with a pressure differential across the magnets of 7kg/cm^2. This flow is maintained by an 80 h.p. pump in the plant room. A reduced flow rate which is often required for experiments at lower field intensities will be maintained by a 20 h.p. pump.

After passing through the magnet and before returning to the reservoir the water passes through a heat exchanger rated to extract 1MW under the usual temperature and humidity conditions. This heat is exchanged to softened water in a secondary circuit and is discharged into the atmosphere via an induced-draft, evaporative heat exchanger sited on the roof just outside the plant room.

Even when experiments require periodic runs at 2MW the average power dissipation seldom exceeds 1MW and the reservoir can always be held to within 10°C of the ambient temperature.

An additional item in the cooling system consists of a 12 h.p. heat pump set which extracts heat from the cooling water reservoir and supplies it to the Mullard Cryomagnetic Laboratory via water pipes in the floors of the building. As the temperature of the source is relatively high this provides an economical system for heating the building for most of the year, and also a method of cooling the tank.

The Performance and Limitations of Low-Level Amplifiers Employing Tunnel Diodes

By B. Easter*, M.Sc., A.M.I.E.E.

The tunnel diode amplifier has potential application as a simple receiver input stage for those systems above 1Gc/s where a moderately low noise factor is acceptable. The performance of this type of amplifier is considered here in relation to the requirements of line-of-sight radio-relay systems. Some detailed consideration is given to the non-linear effects which can arise, including a.m./p.m. conversion. Problems arising in the design of a practical amplifier are discussed, including that of achieving circuit stability. Some details of a design for operation at 2Gc/s are given.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

POTENTIAL applications in the microwave frequency range were widely recognized early in the history of the tunnel diode: in particular, low-level amplification with a moderately low noise factor was seen as an attractive possibility. Diodes suitable for use at microwave frequencies extending up to 10Gc/s and beyond are now generally available in at least experimental quantities. A considerable number of experimental amplifiers have been described in the literature: the order of performance attainable with a tunnel diode amplifier in terms of noise factor and gain-bandwidth is now well established. Information on the application of tunnel diode amplifiers is, however, scanty. It would appear that the present time is a period of assessment, and the likely significance of tunnel diodes in microwave systems should soon become evident. The present article is offered as a contribution in this context.

A representative range for the noise factor attainable with a tunnel diode amplifier is 3 to 6dB, while gain-bandwidth products of the order of 1 000Mc/s may be obtained. Bearing in mind the recent improvements in high frequency transistors, and the almost mandatory use of a circulator for many applications of negative conductance amplifiers, the tunnel diode is seen to be mainly of significance for systems above, say, 1Gc/s. With its attractive simplicity and potential reliability, it could be an optimum choice of input amplifier for systems where the effective source temperature at the receiver is of the order of 300°K, and the advantage to be gained from more complex input stages is thereby limited. Line-of-sight radio-relay systems operating at 2, 4, and 6 to 7Gc/s are a pertinent example of some importance, and the tunnel diode amplifier will be considered here in relation to the requirements of these systems. Among these requirements, special attention must be given to the demands of reliable, unattended operation, and low signal distortion.

Characteristics of Available Tunnel Diodes: Limits of Amplifier Performance

The general form of the conduction characteristic of the tunnel diode is well known. For low-level amplification, interest is centred on the negative conductance region which extends from the current peak, at a bias voltage of 50 to 65mV for germanium, to the current minimum, or valley, at a bias of about 300mV. For a given material, the voltage scale of the characteristic varies only slightly, while the current range is proportional to the peak current of the device. The latter can be well controlled to a tolerance of a few per cent in manufacture; thus the negative conductance characteristic is also relatively well controlled. Fig. 1 shows the small signal equivalent circuit for operating bias in the neighbourhood of maximum negative conductance. Parameter values are given for two representative classes of diode; the peak current is 2mA nominal in both cases, to give parameters broadly compatible with orthodox circuit impedances. The elements R_n, R_s and C_j are associated with the diode junction

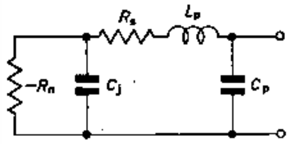

Fig. 1. Equivalent circuit and typical element values

	(a)	(b)
Operating frequency	1·5-4Gc/s	4-8Gc/s
Cut-off frequency	6-12Gc/s	10-20Gc/s
R_n	50Ω	50Ω
C_j	1·5-4pF	0·8-2pF
R_s	0·5-1·5Ω	0·6-2Ω
L_p	300pH	100pH
C_p	1pF	0·3pF
Pill dimensions	0·100in × 0·050in	0·050in × 0·025in

region, and may be regarded as lumped elements in the frequency range considered: the remaining elements L_p and C_p are associated with the encapsulation and must be interpreted with due regard for the manner of mounting the diode in the external circuit. The 'pill' construction commonly adopted is the result of an effort to reduce these latter elements, particularly the inductance.

Two characteristic frequencies typify the high frequency limitations of a diode. The resistive cut-off frequency given by:

$$f_c = \frac{1}{2\pi C_j \sqrt{(R_s R_n)}} \sqrt{(1 - (R_s/R_n))}$$

gives an upper limit to the active range: operation will normally be restricted to one-third, or possibly one-half of this figure. The self-resonant frequency

$$f_{xo} = \frac{1}{2\pi \sqrt{(L_p C_j)}} \sqrt{(1 - (L_p/C_j R_n^2))}$$

does not have such sharp significance, but is useful as a guide to the circuit designer.

In terms of the above parameters, limiting values can be given for the noise factor and bandwidth, assuming a

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singly resonant amplifier coupled to a circulator:

$$F = \frac{1 + 20\, I_o R_n}{(1 - (f/f_o)^2)\,(1 - R_s/R_n)}$$

where I_o is the d.c. bias

$$B \simeq \frac{1}{(G^{\frac{1}{2}} - 1)} \cdot \frac{1}{\pi R_n\,(C_j + C_p)}$$

where G is the power gain.

The expression for noise factor results from simplifying assumptions attributing full shot noise to the net bias current, and thermal noise to the spreading resistance: it has been found empirically that the expression is of adequate accuracy in the normal bias range. For a germanium diode of high quality, the term $20\, I_o R_n$ has a value in the range 1·2 to 1·3.

Alternative materials have been considered, and devices fabricated in gallium arsenide and gallium antimonide can be obtained. Compared to germanium, gallium arsenide offers an improvement of about 6dB in power handling capacity for a given diode impedance, but with an increase of noise factor of some 2dB. Gallium antimonide yields an improved noise factor at the price of significantly lower signal handling capacity and greater sensitivity to temperature variations. It is thought that germanium will probably remain the common choice for amplifier diodes.

With operation at a fixed bias, the negative conductance of a germanium diode is associated with a negative temperature coefficient which is typically 0·25 per cent/°C in the range 0°C to 50°C. This variation is of the same order as that experienced for the characteristic impedance for some junction circulators: it is possible to arrange a degree of compensation of the two effects. Alternatively variation of the bias with temperature can be used to realize satisfactory performance over a range of temperature.

Non-linear Effects

Inspection of the characteristic of a tunnel diode immediately emphasizes the low power handling capacity of the device; thus, for a typical 2mA amplifier diode, the power drawn from the bias supply is normally about 120μW, and significant non-linearity is to be expected for amplifier input levels of the order of 1μW for useful values of gain. In particular, the margin relative to the received signal under low-fading conditions in a representative microwave relay system is likely to be small.

The major cause of non-linearity is, of course, associated with the shape of the conduction characteristic, but the voltage dependent capacitance of the junction is also significant. A rigorous discussion of the non-linear behaviour will not be attempted: however a useful first order estimation of the performance may be obtained by distinguishing a number of distortion mechanisms.

For the present discussion an idealized amplifier form will be assumed, with the diode connected in a shunt resonant circuit and coupled to a circulator: resistive losses due to the spreading resistance or the circuit dissipation will be neglected. Corrections to take account of these losses will be discussed separately.

Amplitude Compression due to the Non-linearity of the Characteristic

Relative to a given operating point, the conduction characteristic can be represented by a power series:

$$i = -G_n v + G_2 v^2 + G_3 v^3 + \ldots$$

In particular, in the neighbourhood of the point of maximum conductance, the characteristic is well represented by:

$$i = -G_n v + G_3 v^3$$

The effective fundamental conductance for a sinusoidal voltage of peak amplitude V may easily be shown to be given by:

$$-G_n' = -[G_n - \tfrac{3}{4}\, G_3 V^2]$$

The reduction of gain for a given input level can be estimated.

For a typical 2mA diode, biased for maximum conductance,

$$G_n \simeq 0{\cdot}02\text{mho};\ G_3 \simeq 3\text{A/V}^3$$

Taking these values the input signal level for 1dB loss of gain is estimated to be −36·5dBm for a low level gain of 14dB; −45·5dBm for 20dB gain. These figures illustrate

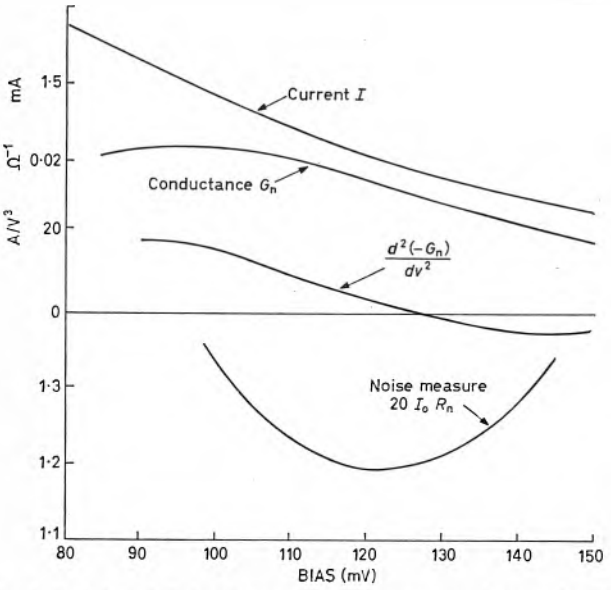

Fig. 2. Variation of diodes characteristics with respect to the operating bia

the enhanced dependence of the distortion on the gain which is to be expected for a negative conductanc amplifier.

In practice, a better choice of operating point can b made as illustrated by Fig. 2. The variation wit bias voltage is shown for the diode current, the nois measure $20\, I_o R_n$, the slope G_n and the second derivativ of the slope which is proportional to G_3. Bias voltage in the region of minimum noise measure yield values, o G_3 approaching zero; an appreciable improvement of th overload characteristic is to be expected, although highe order terms will become significant. Empirical overloa characteristics of a practical amplifier will be given in later section.

The above considerations indicate the desirability of careful choice of the operating gain: this is also relevan to the need for long-term stability and acceptable tem perature variations. A figure in the region of 14dB i likely to represent a satisfactory compromise.

Amplitude to Phase Conversion

While the 1dB compression point is widely taken t represent the signal handling capacity of a low-leve amplifier, it must be emphasized that amplitude compre sion is not of direct importance in an f.m. system amplitude to phase conversion is of greater significanc

Several sources of amplitude to phase conversion ca be expected:

(a) The decrease of the effective diode conductance at high input levels, as discussed above, will also be directly associated with an amplitude dependent phase shift if the input frequency departs from the resonant frequency of the circuit: the load impedance may have a reactive component due to temperature variations in the circulator, for example. If the net susceptance is equivalent to a load v.s.w.r. of 1·1, then for a gain of 14dB, 1dB compression will be associated with a phase shift of 1·5° relative to low level transmission.

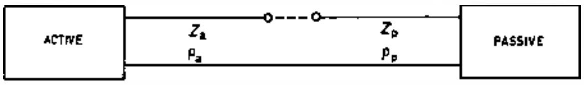

Fig. 3. The essence of the stability problem

(b) An increase in the average diode bias current will occur at high input levels: this can be related to the even order terms of the characteristic expression as considered above. Variation of the bias voltage will occur, dependent on the effective bias supply impedance: modifications of the effects already described will result. In addition, due to the voltage dependent junction capacitance, i.e. the depletion layer capacitance of an abrupt junction, a further phase shift will be introduced. These effects can evidently be controlled by appropriate design of the bias supply to provide a low impedance, with particular regard to the modulation frequency range of the system.

(c) If constant bias voltage is assumed, the mean capacitance of the junction will be increased when the non-linear capacitance is 'pumped' by a relatively large signal. This effect typically results in 1° phase shift (relative to low levels) at 1μW input and 14dB gain.

The relative phase shifts discussed above will all be roportional to a first order, to the square of the signal oltage magnitude. It would appear that an a.m./p.m. conersion factor of up to $\frac{1}{2}$°/dB is to be expected at the 1dB ompression point, for a gain of 14dB. It is to be noted at, for a given input level, the relative phase shift and e conversion factor will both be proportional to the cube f the voltage gain ratio.

NFLUENCE OF HARMONIC CURRENTS

Finally, a more complex distortion mechanism must be nsidered. At relatively high signal levels, a significant ow of harmonic current will occur at the junction. In articular, for an operating point chosen for optimum oise factor, there will be a significant magnitude of cond harmonic current. A harmonic voltage will be nerated, depending on the total admittance appearing the junction at the second harmonic frequency. This turn will interact with the fundamental voltage via the n-linear characteristic to generate a distortion comnent at fundamental frequency. Space does not permit detailed analysis of this effect. In brief, it is found that e effective fundamental admittance of the diode includes component depending, in phase and magnitude, on the ase and magnitude of the second harmonic admittance, d proportional to the square of the fundamental voltage agnitude. There will occur, therefore, amplitude variams and phase shifts having a similar variation with input el to the effects considered earlier. In principle, distorn due to this effect could be severe; however this would only seem to occur when the circuit is rather close to oscillation at the harmonic frequency. Proper design should ensure that there is an ample margin of stability, particularly at the second harmonic frequency.

EFFECTS OF DISSIPATION

The above discussion of non-linearity neglected the influence of the spreading resistance R_s and circuit dissipation. It is often possible to represent, with some approximation, all the losses by a single shunt conductance, G_d referred to the diode junction: it is necessary to assume $R_s \ll R_n$, valid for a good quality amplifier diode.

In these terms, the noise factor is given by

$$F \simeq \frac{1 + 20\, I_o R_n}{1 - G_d R_n}$$

With regard to non-linearity, it is significant that, relative to the idealized loss-free amplifier, the sensitivity of the gain to variations of diode conductance, and the signal power at the diode junction are both increased by the factor

$$\frac{1}{(1 - G_d R_n)}$$

Stability

The attainment of satisfactory circuit stability can impose serious difficulties and limitations in the realization of a negative conductance amplifier. The availability of low-loss, well matched circulators solves the central problem of utilizing a one-port amplifier, but the restricted bandwidth of practical circulators, compared with the wide active frequency range of the tunnel diode, is often associated with possible modes of oscillation outside the operating frequency range. The design problem is that of suitably controlling the circuit impedance presented to the diode without introducing significant circuit dissipation in the operating frequency range: the whole frequency range up to the resistive cut-off of the diode must be considered. A full development of the subject is not possible here; however a brief introduction to the form of useful stability criteria can be given.

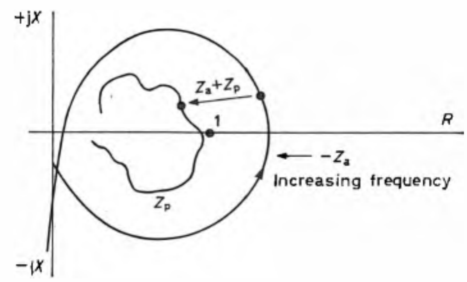

Fig. 4. Illustration of a criteria for stability

Standard texts on circuit theory, such as that of Bode[1] give consideration to the stability of active networks, but the derivation of useful criteria for the present situation is not obvious. The most useful contribution would appear to be that of Henoch and Kvaerna[2].

As illustrated in Fig. 3, the overall circuit may be considered to be arbitrarily divided into two series connected one-ports: an active impedance Z_a including the tunnel diode, and a passive impedance Z_p including the circulator, input and output circuits. Typically, parts of the latter circuit will only be loosely defined over a large portion of the frequency range. The problem to be solved is commonly that of determining the modifications of Z_a or Z_p necessary to achieve stable operation.

The impedance sum $[Z_a + Z_p]$ may be considered as a function of the complex frequency variable $p = j\omega$. For stable operation $[Z_a + Z_p]$ must be short-circuit stable: there can be no zeros of positive real part. Poles of positive real part can only be contributed by Z_a. It follows, by one of the theorems of complex variable theory, that for a given Z_a, stability is associated with a definite number of encirclements of the origin of the complex

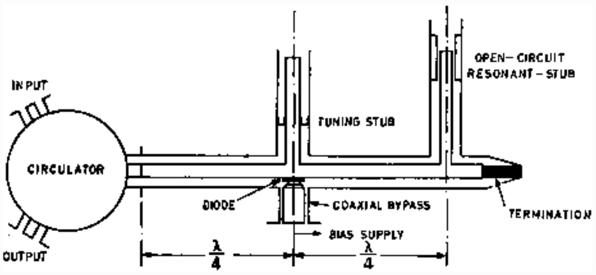

Fig. 5. Schematic of a 2Gc/s amplifier

Gain 14dB; Noise factor 4·4dB; Input level −30dBm; Bandwidth 200Mc/s (1dB compression)

impedance plane as $[Z_a + Z_p]$ is plotted over an infinite range of negative and positive real frequency. Thus, if there is detailed knowledge of Z_a, or alternatively, if Z_a is known to be stable with a broad band matched load, for example, then the stability for an arbitrary load circuit Z_p can be determined.

Equivalent relations in terms of the admittances or the reflection coefficients can be obtained.

In particular, if the amplifier is stable with a broad band matched termination, a useful criterion for stability in terms of reflection coefficients is given by

$$|\rho_a \cdot \rho_p| < 1$$

In theory, however, this criterion is unnecessarily restrictive.

A more practically useful interpretation of the above theory is illustrated by Fig. 4. The plot of $[-Z_a]$ for a simple amplifier is represented on the normalized complex impedance plane: values for positive frequency only are given for the sake of clarity. If now Z_p is plotted on the same diagram, $[Z_a + Z_p]$ is given by the vector difference of the two plots. The behaviour of $[Z_a + Z_p]$ as the nature of Z_p is modified can often be directly visualized with this presentation, giving useful indications of the allowable behaviour of Z_p. In the simple case presented, stable operation will occur if Z_p is the unit resistance, and also if Z_p is constrained to remain within the closed loop of the $[-Z_a]$ plot: Z_p can, in addition be allowed to vary outside the closed loop providing the number of mutual encirclements of the two plots is not thereby altered. This interpretation can equally well be performed on the reflection coefficient plane, or Smith chart.

Amplifier Construction and Performance

For applications where the bandwidth is adequate, the simple singly resonant circuit arrangement is attractive. For the lower frequencies, where the influence of the effective series inductance is not too great, a shunt resonant circuit design may be realized which is reasonably close to the idealized form. A satisfactory margin of stability can be achieved by the introduction of a selective circuit which provides a positive conductance at the diode terminals outside the operating frequency range. The arrangement is made clear by the simplified schematic of Fig. 5. Both coaxial and strip line configurations have been used in amplifiers of this form: in either case the cross-section of the circuit must be compatible with the diode dimensions if the effective diode inductance is not to be unduly increased.

The bias circuit of the diode demands careful attention: in particular it must have a resistive component of impedance at the operating frequency which is low compared to the spreading resistance of the diode, but must be free of resonant effects at higher frequency which could cause instability. Means must be included for exerting a controlled contact pressure on the contact faces of the diode ‘ pill ’.

An established design at 2 000Mc/s of the general form of Fig. 5 has the following performance:

Gain	14dB
Noise factor	4·4dB
Bandwidth (−3dB)	200Mc/s
Input level (1dB compression)	−30dBm

The noise factor is in excellent agreement with theory. The bandwidth in this case is appreciably decreased by the influence of the stabilizing circuit: if necessary an increase could be obtained by replacing the single stub by a more complex band-stop filter.

The performance of the circuit in the overload region is shown in Fig. 6, where the variations of gain and relative phase shift are given for a range of input level. Overall noise measurements on a medium capacity microwave system have confirmed the performance of the amplifier, and in particular, the absence of significant distortion for inputs below the 1dB compression level.

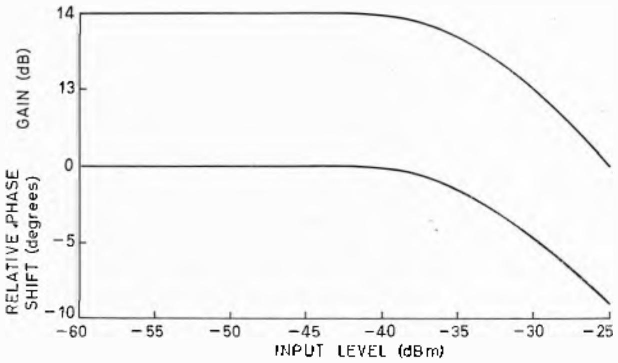

Fig. 6. Overload characteristics of a 2Gc/s amplifier

For operating frequencies above, say, 4Gc/s, increasin difficulty besets the form of design described above. Unles it is possible to depart from the conventional circuit con struction using a separately encapsulated diode, it is neces sary to accept the fact that the mounted diode i approaching self resonance at the operating frequency. I the diode is used as a series resonant element the problen arises of appropriately loading a low negative resistance o a few ohms in magnitude: furthermore, the latter will b dependent on the junction capacitance, which has a wid spread in manufacture. The employment of a multipl resonant coupled circuit offers considerable advantag including enhanced operating bandwidth.

REFERENCES

1. Bode, H. W. Network Analysis and Feedback Amplifier Design. Sectio 3.2. (D. Van Nostrand Co., New York).
2. Henoch, B., Kvaerna, Y. Stability Criteria for Tunnel Diode Amplifier *IRE Trans. Microwave Theory and Tech. MTT-10,* 397 (Sept. 1962).

A Simple Transistor Large Signal Current Gain Measuring Circuit with Digital Presentation

By S. L. Hurst*, B.Sc.(Eng.), A.M.I.E.E., A.M.I.E.R.E.

A simple d.c. test circuit is described, which with the aid of any conventional digital frequency meter enables the value of the common emitter large signal current gain of transistors at various current levels to be directly displayed. The accuracy of measurement is comparable to normal commercial d.c. transistor testers, but with the facility of easy and unambiguous read-out. Common emitter leakage currents may also be measured, and the matching of the current gain performance of transistors may be rapidly made with high matching accuracy.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

RECENTLY published articles[1,2] have detailed the use of digital frequency meters for the measurement of parameters other than frequency and time, such arrangements involving the use of some simple test circuit external to the digital meter to convert the unknown parameter value into an appropriate number of cycles or time interval, etc. The increasing availability of standard digital frequency meters in many organizations has added to the attractiveness of such measurements.

A parameter frequently required to be measured by all purchasers and users of transistors is the common emitter large signal forward current transfer ratio or 'current gain' h_{FE}, and thus the possibility of a test circuit plus digital frequency meter to display the value of this parameter would appear to be a very worth while proposition. ſo be a widely valuable proposition however, the necessary ancillary test circuit should not be complex and expensive; extreme accuracy should therefore be sacrificed for economy if necessary.

Considerations of h_{FE} Measurements

The precise definition of h_{FE} is $[(I_C - I_{CEO})/I_B]$, all currents being measured at some constant collector to emitter voltage V_{CE}, and at some specified value of collector current I_C. A more practical definition, and one used by many semiconductor authorities, is to deliberately neglect the I_{CEO} leakage term and define h_{FE} as $\left.\frac{I_C}{I_B}\right|_{V_{CE}}$, at the particular specified collector current. This latter definition has the anomaly of making the apparent value of h_{FE} increase at very low current levels as the small value I_{CEO} term becomes increasingly dominant and no longer negligible, but in general the transistor should never be operated at such impractically low current levels.

Thus accepting the latter definition for h_{FE}, to obtain a value for this parameter necessitates pre-setting the collector current I_C and the collector to emitter voltage V_{CE} to their desired values, and then adjusting the input current I_B in order to simultaneously achieve both these preset requirements. I_B is thus fundamentally the variable quantity in the desired expression. To automatically realize this balance of conditions in practice, however, immediately necessitates a relatively complex test circuit arrangement, such as shown in Fig. 1 for example.

In this arrangement I_C may be preset to the desired value by adjustment of R_C, this current being substantially independent of V_{CE} provided supply voltage V_{CC} is large. The base input current is now governed by the high gain differential amplifier, I_B being increased as necessary until the desired value of V_{CE} results. By now measuring the voltage drop across R_B, this voltage may be suitably scaled so as to indicate directly the value of h_{FE} for the transistor under test.

The disadvantage of this test circuit is the stringent requirements for the d.c. coupled differential amplifier,

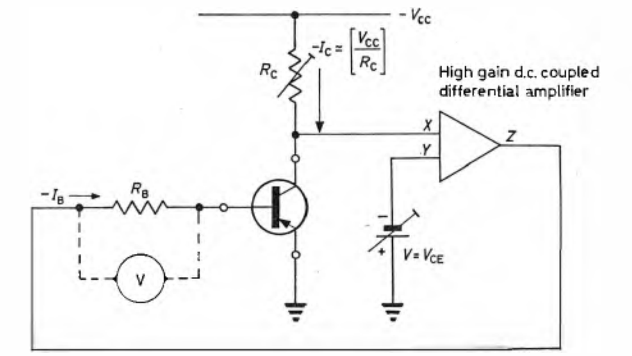

Fig. 1. Arrangement of test circuit for determination of h_{FE}, with I_C and V_{CE} preset

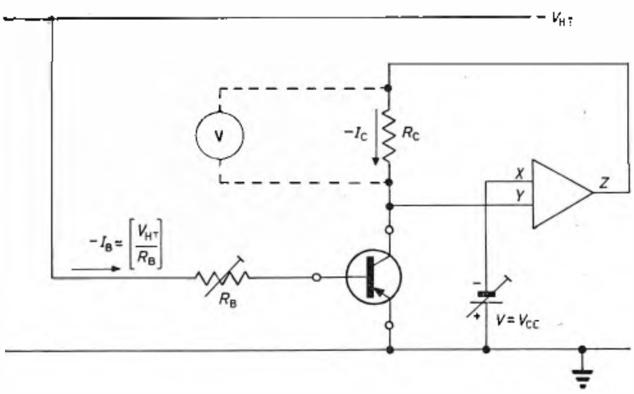

Fig. 2. Arrangement of test circuit for determination of h_{FE}, with I_B and V_{CE} preset

which must have a very high gain and a high input impedance, and also the inconvenience in scaling the voltage drop across R_B to read h_{FE} directly, due to the inverse relationship between these two factors. In all, therefore, although this is the basic form of some very precise h_{FE} test circuits, it does not lend itself to any simple economical realization.

Some simplification of the problem is achieved if the conditions for the measurement of h_{FE} are relaxed from

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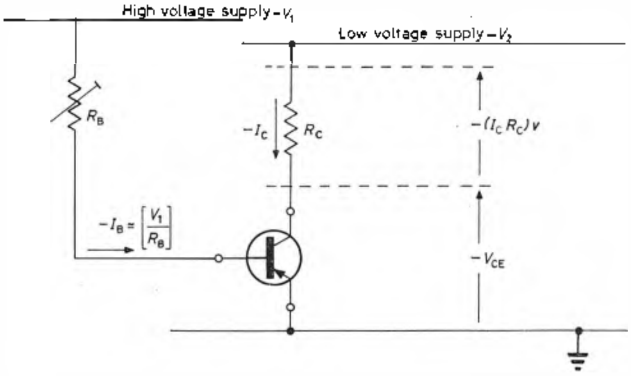


Fig. 3. Arrangement of test circuit for determination of h_{FE}, with I_B preset, V_{CE} variable

I_C, V_{CE} constant, to I_B, V_{CE} constant, thus making I_C the variable quantity when under test. This is not usually such a directly useful determination of h_{FE} for the circuit designer, as in practice maximum I_C is fixed, and I_B adjusted as necessary. However, these differences in determination do not in general give rise to any great inconvenience, and h_{FE} measurements with I_B rather than I_C preset and constant are the rule rather than the exception in the majority of commercial test circuits. A test circuit for such a measurement, holding V_{CE} to some preset value may therefore be as shown in Fig. 2. Variations on this circuit may also be encountered, for example presetting I_B and holding V_{CB} zero instead of V_{CE} to some small finite value.

Stringent requirements still exist for the differential amplifier, but the scaling is now simplified as the voltage drop across R_C is now directly proportional to h_{FE}. However, the continuing necessity for a good d.c. coupled differential amplifier, with the attendant d.c. stabilization problems, still makes such a test circuit a relatively expensive proposition.

Relaxation of the precise requirement to hold V_{CE} or V_{CB} constant while measuring h_{FE} immediately enables the differential amplifier to be deleted, and this therefore points the way to an economical solution. Such a solution is of course the basis of all small portable commercial transistor testers currently available, the basic test circuit being as shown in Fig. 3. Here I_B is preset to an approximately constant value by adjustment of R_B connected to the relatively high voltage V_1, while V_{CE} perforce varies due to the varying voltage drop in R_C, which is directly proportional to the resulting transistor collector current I_C. This voltage variation may be minimized by choice of R_C such that the voltage drop in R_C is not greater than, say, 1V, which if V_2 is, say, 3V means that V_{CE} may lie between 2 and 3V.

Thus with this simple circuit, the value of h_{FE} of the transistor under test may be obtained by scaling either the resultant current flowing through R_C or the voltage drop developed by this current across R_C, both being directly proportional to the value of h_{FE}. In most commercial transistor testers, it is the current I_C passed through an appropriately scaled ammeter which is employed to read h_{FE}, but in the following test circuit the voltage drop produced by I_C will be used.

Method of Conversion from Voltage to Frequency

Having decided upon the basic method of measuring h_{FE}, the problem of conversion of the d.c. output voltage of this circuit to some proportional or linearly related frequency becomes the major part of the remaining test circuit.

With simplicity and economy still in mind, one of the simplest frequency sources is the symmetrical astable transistor multivibrator circuit. While this circuit is generally regarded and used as a fixed or preset frequency source, its great attraction being that its frequency is substantially independent of d.c. supply voltage, it may nevertheless be very simply modified to produce a frequency that varies with voltage[3]. Two simple related modifications are available, the first of which produces a wide percentage change of frequency but with a relatively low input resistance to the adjusting d.c. voltage, the second producing a small percentage change of frequency but with a higher value input resistance. Both methods can be arranged to produce a substantially linear change of frequency with the variable d.c. voltage.

For this particular application the wide percentage change of frequency circuit is most advantageous. The circuit diagram for this arrangement is shown in Fig. 4(a), with typical frequency/variable d.c. voltage relationships in Fig. 4(b). The latter indicates that a choice of both the

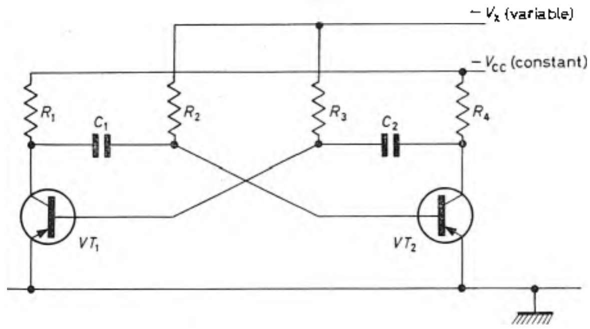
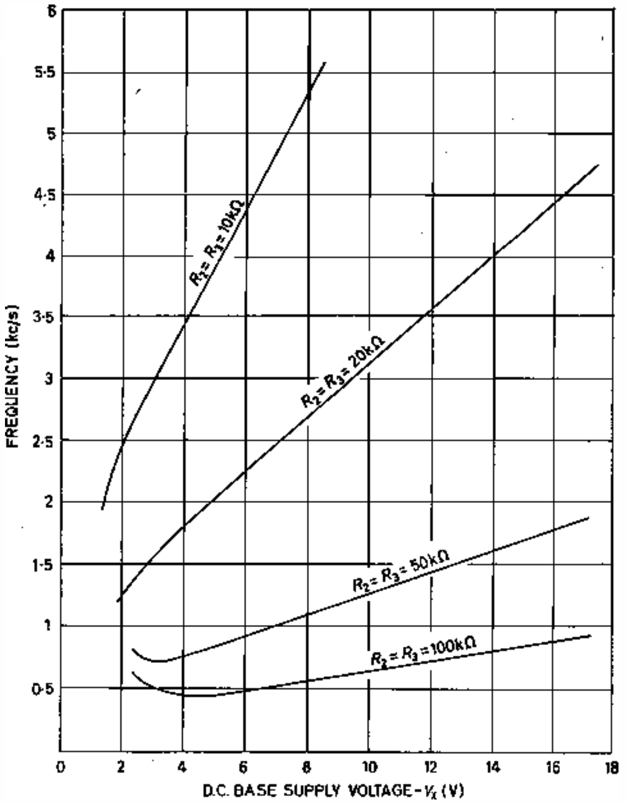


Fig. 4(a). 'Wide range' variable frequency astable multivibrator circuit

(b). Frequency characteristics of circuit of Fig. 4(a) with $V_{CC} = -12$V, $C_1 = C_2 = 0{\cdot}01\mu$F, $R_1 = R_4 = 2{\cdot}2$kΩ



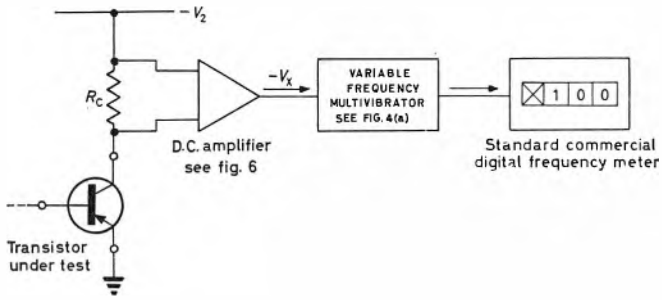


Fig. 5. Arrangement of complete test circuit and read-out

actual frequency and also the rate of change of frequency $[\Delta f/\Delta V_X]$ is available.

Choosing $R_1 = R_4 = 2{\cdot}2\text{k}\Omega$, $C_1 = C_2 = 0{\cdot}01\mu\text{F}$, $R_2 = R_3 = 50\text{k}\Omega$, then with $V_{CC} = -12\text{V}$ and $V_X = -7\text{V}$ a frequency of oscillation of 1kc/s results, with $[\Delta f/\Delta V_X] = 87{\cdot}5$c/s per volt $= 1{\cdot}143$V per 100c/s. Suppose now that the value of collector resistor R_C in the transistor test circuit of Fig. 3 is chosen such that there is a voltage of 0·5V developed across it when the transistor under test has a value of $h_{FE} = 100$, the interposing of a d.c. amplifier between R_C and the variable frequency multivibrator with a quiescent output voltage of -7V and with a voltage gain of 2·286, will result in the multivibrator frequency being modified from 1kc/s to 1·1kc/s when the transistor is under test.

Thus ignoring or blanking off the thousands decade indication of a digital frequency meter that is connected to record the multivibrator frequency, a presentation directly equal to the value of h_{FE} of the transistor under test will be obtained. The block diagram of the complete test circuit thus becomes as shown in Fig. 5, with the d.c. amplifier plus multivibrator circuit giving an overall performance of:

$$\left[\frac{\text{Amplifier input (mV)}}{5}\right] = \begin{array}{c}\text{Change of multivibrator}\\ \text{frequency (c/s)}\end{array}$$

The D.C. Amplifier

The requirements for the d.c. amplifier are

(1) That its quiescent output d.c. voltage must be adjustable to give the required 1kc/s 'zero' output frequency with zero voltage input fed across R_C.

(2) That its gain must be adjustable to give the desired rate of change of frequency with input voltage.

(3) That its input/output voltage must be in phase. In addition its output impedance must be sufficiently low as to be able to adequately supply the requirements of the multivibrator circuit.

A great degree of complexity may very rapidly be involved if a highly stable d.c. amplifier circuit is attempted. Similarly any attempts to completely isolate the effects of the preset gain control and the preset quiescent output voltage control so that each may be separately adjusted without influencing the other, immediately leads to complications and the proliferation of amplifier stages.

The simplest solution that can be proposed is as shown in Fig. 6, where no attempt has been made to achieve high d.c. stabilization, or to isolate the effects of the two preset controls one from another. The small variation of output d.c. level is obtained by adjustment of RV_1, which permits alteration of the effective d.c. supply voltage for VT_4, while RV_2 allows alteration of the voltage gain of the circuit, though inevitably also altering the final output d.c. level as well. It is, however, found that these two interacting controls can in practice be set very rapidly to their required values, and the considerable complication of a theoretically more satisfactory circuit appears unnecessary.

The input voltage to the base of VT_3 of this amplifier can most conveniently be made by making R_6 of Fig. 6 be the collector resistor R_C of the transistor under test. The d.c. voltage supplies V_1 and V_2 of the unknown transistor must however now be completely isolated from supply V_{CC} of the amplifier and multivibrator circuits, but as these are inevitably different value voltage supplies this is no undue inconvenience.

The Complete Test Circuit

Fig. 7 shows the complete test circuit. R_6 will be seen to finally be composed of a chain of resistors, the appropriate total value as seen by the unknown transistor current I_C being selected as required to match the preset base input current setting, as will be detailed below. Resistors R_2 and R_3 have been made standard values, which together with the normal tolerances on all components means that the quiescent output voltage and the voltage gain of the d.c. amplifier are slightly different from the precise values quoted above.

The resistor values necessary for R_6 and the transistor base input resistor R_B are best illustrated by means of the tabulation in Table 1. The values shown for R_B are assuming a stabilized voltage supply $V_1 = 87$V, with a negligible transistor base-to-emitter voltage in comparison with this voltage.

TABLE 1

Corresponding Resistance values for R_B and R_6

I_B	REQUIRED VALUE OF R_B ($\pm1\%$)	POSSIBLE VALUES OF h_{FE}	RESULTANT CURRENT I_C	REQUIRED COLLECTOR RESISTANCE R_C	RESULTANT VOLTAGE DROP $R_C I_C$	RESULTANT Δf ON DIGITAL METER (c/s)	REQUIRED VALUE OF R_6 ($1\pm\%$)
3mA	29kΩ, $=R_B(f)$	1 50 200	3mA 150mA 600mA	1·67Ω	5mV 250mV 1·0 V	1 50 200	$R_6(f)$ $=1{\cdot}67\Omega$
1mA	87kΩ, $=R_B(e)$	1 50 200	1mA 50mA 200mA	5Ω	5mV 250mV 1·0 V	1 50 250	$R_6(e)$ $-3{\cdot}33\Omega$
300μA	290kΩ, $=R_B(d)$	1 50 200	300μA 15mA 60mA	16·7Ω	5mV 250mV 1·0 V	1 50 200	$R_6(d)$ $=11{\cdot}7\Omega$
100μA	870kΩ, $=R_B(c)$	1 50 200	100μA 5mA 20mA	50Ω	5mV 250mV 1·0 V	1 50 200	$R_6(c)$ $=33{\cdot}3\Omega$
30μA	2.9MΩ, $=R_B(b)$	1 50 200	30μA 1·5mA 6mA	167Ω	5mV 250mV 1·0 V	1 50 200	$R_6(b)$ $=117\Omega$
0	O.C.	—	$I_{CBO}=$ say 50μA	5000Ω	250mV	50	$R_6(a)$ $=4833\Omega$

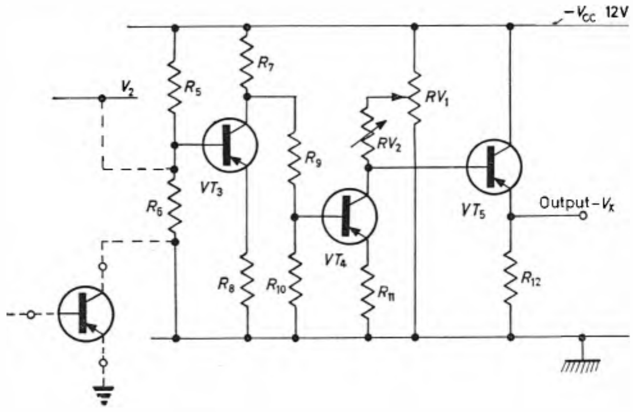


Fig. 6. Simple d.c. amplifier circuit, with output $-V_x$

From Table 1 it will be noted that the hundreds, tens, and units digits of the digital frequency meter will always display directly the value of h_{FE} of the transistor under test. The final row of figures indicates that the collector-to-emitter leakage current I_{CEO} may also readily and directly be displayed by this test circuit as well as h_{FE}, by open-circuiting the base circuit and making R_6 an appropriately larger value.

The necessary presetting of the output d.c. level and the voltage gain of the amplifier, governed by the potentiometers RV_1 and RV_2 of the amplifier circuit, is facilitated by the provision of the three-position switch S_2, positions (b) and (c) of which being the 'calibrate' positions. In position (b), all inputs to the R_6 resistor chain are open-circuited, and the amplifier presets should be adjusted to give a final multivibrator output frequency of 1kc/s. In position (c), a preset amount of current is passed through R_6 to form a fixed input signal of 200μA, and the controls adjusted to now give an output frequency of 1·2kc/s.

Control RV_2 has the dominant effect on the gain of the amplifier circuit, and thus can rapidly be set to give a difference frequency of 200c/s between the above two 'calibrate' positions, following which RV_1 can be adjusted to give the correct actual frequencies of 1kc/s and 1·2kc/s. Once set, only very minor re-adjustments of these controls become necessary on subsequent occasions of using the test circuits.

It is, of course, necessary to ensure that the bias current for transistor VT_3 flowing through R_5 is not shunted by the variable impedance circuit of the transistor under test, and conversely that no alternative path for the latter collector current exists except through the common resistor chain R_6. The complete isolation of these two circuits is achieved by the diodes MR_1, MR_2, and MR_3 shown in Fig. 7, which ensure that this requirement is maintained.

Accuracy of Measurement

It is of academic interest only to record the linearity and accuracy of each individual part of the test circuit. Instead overall accuracy becomes a far more meaningful factor.

However a tabulation to indicate the accuracy between current fed into the resistor chain R_6, = the 'unknown transistor' collector current I_C, and the resultant digital read-out is relevant, such information being given by the typical figures shown in Table 2. These figures are obtained with ± 1 per cent tolerance resistors in the R_6 resistor chain, and with the correct presetting of the 1kc/s and 1·2kc/s spot frequency outputs.

As with all digital read-outs, the least significant digit may vary by one during any specific count. This is a high percentage error if the test circuit is measuring a low value of h_{FE}, say 10, or a small leakage current I_{CEO}, say

Fig. 7. The complete test circuit, with no compensation for I_{CEO} leakage current

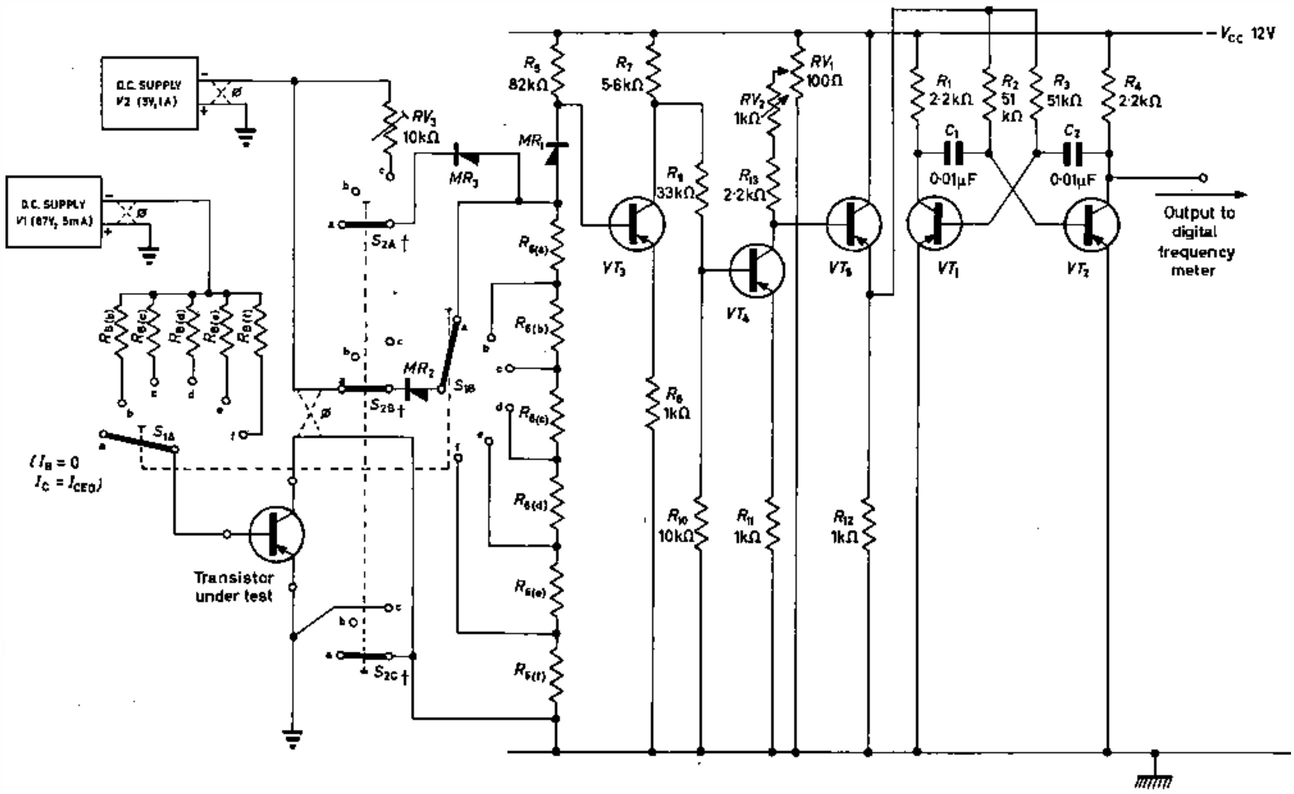


TABLE 2

Typical Results Given by Amplifier/Variable Speed Multivibrator/Digital Frequency Meter Assembly

INPUT CURRENT I_C THROUGH R_6	TOTAL SELECTED VALUE OF R_6 (Ω)	THEORETICAL DIGITAL METER READING (c/s)	ACTUAL DIGITAL METER READING (c/s)
0		1000	1000
25μA		1025	1026
50μA		1050	1051
100μA	5000	1100	1101
150μA		1150	1151
200μA		1200	1200
2·5mA		1025	1026
5·0mA		1050	1052
10·0mA	50	1100	1102
15·0mA		1150	1152
20·0mA		1200	1201
25mA		1025	1026
50mA		1050	1052
100mA	5·0	1100	1102
150mA		1150	1152
200mA		1200	1201

10μA, and an alternative scaling of the effective resistor value R_6 to give a $\times$ 10 factor on the read-out could be provided to minimize this feature if felt necessary.

An additional error that may occur over and above these previous factors when measuring h_{FE}, is the tolerance on the base input current I_B to the transistor under test. Taking this latter factor into account, and allowing a maximum tolerance of ± 2 per cent, the overall accuracy of measurement of h_{FE} will under the worst conditions be within 5 per cent $\pm$ 1 count. For I_{CEO} measurements, where the tolerance on I_B is immaterial, then ± 3 per cent $\pm$ 1 count may be realized.

Back-off of I_{CEO} Leakage Current

If it is desired to measure h_{FE} taking into account the collector to emitter leakage current, thus measuring h_{FE} as $[(I_C - I_{CEO})/I_B]$ with I_B preset, V_{CE} variable, the additional back-off circuit shown in Fig. 8 may be incorporated. When now measuring I_{CEO}, this additional circuit requires to be broken, and a push-button PB_1 has to be incorporated for this purpose.

To adjust RV_4 to just back off the leakage current, the S_1 selector switch should be on its 'Measure I_{CEO}' position, push-button PB_1 not depressed, and RV_4 then adjusted until zero leakage current flows through the resistor chain R_6, as denoted by the 'zero' 1 000c/s frequency reading on the digital frequency meter.

Although the vast majority of small d.c. transistor testers commercially available do incorporate some form of I_{CEO} back-off when undertaking h_{FE} measurements, it is very debatable whether h_{FE} measured in this way is any more useful to the user of the transistor than h_{FE} measured without I_{CEO} back-off as merely $[I_C/I_B]$. Intelligent appreciation of the magnitude and effect of the I_{CEO} leakage term must, however, be born in mind with the latter preferable definition.

Conclusions

The test circuit described provides a means of h_{FE} measurement which gives an accuracy comparable with most normal commercial d.c. transistor testers, but with the advantage of unambiguous digital read-out. The resolution time is the resolution time of the frequency counter employed, which invariably for the frequency range under consideration will be 1sec. This time is possibly less than the time taken to correctly read off the usual meter indication reading of a commercial tester.

The ease and relative rapidity with which measurements of h_{FE} can be made at several values of base current input, enables a quick assessment to be made of the approximate value of the small signal common emitter current gain h_{fe} at the various current levels. Without graphical plotting, such a determination must be approximate, but probably sufficient accuracy for many simple amplifier applications would be realized.

As with all forms of d.c. transistor testers, care must be taken with this test circuit also not to overload the transistor under test. This problem of transistor dissipation while under test can only be satisfactorily resolved by resorting to short duration pulse testing techniques, but such techniques are inevitably vastly too complex for any simple and economical realization.

Finally, it may be noted that the essential action

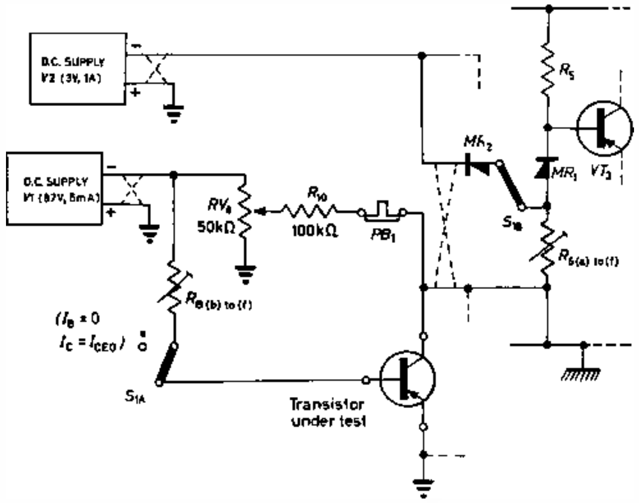


Fig. 8. Addition to test circuit to provide back-off I_{CEO} leakage current

of this test circuit is an analogue to digital current-to-frequency conversion. For this particular application, no special measures to ensure extreme linearity of the conversion were taken or were justified, but even with the simple circuit disclosed, acceptable linearity is revealed in the tabulation of Table 2. This indicates that if greater emphasis is placed on the design of a very stable linear d.c. amplifier, and optimization of the voltage and component values of the variable speed multivibrator circuit is undertaken, a very linear current-to-frequency conversion circuit over wide ranges of current may be possible.

Acknowledgments

Thanks are gratefully acknowledged to Mr. J. Pearce, Senior Engineer, Digitizer Techniques Ltd., who first suggested the idea and attractiveness of measuring transistor parameters with a digital read-out, and also to Professor A. J. Eales, Head of Department of Electrical Engineering, Bristol College of Science & Technology, for facilities made available in pursuit of this circuit investigation.

REFERENCES

1. SPENCER, A. J. Measurement of Phase Angle using a Counter. *Marconi Instrum.* *8*, 63 (Sept. 1963).
2. HURST, S. L. A Simple Capacitance Measuring Circuit with Digital Presentation. *Electronic Engng.* *36*, 596 (1964).
3. HURST, S. L., SEAL, C. L. Frequency Control of Astable Transistor Multivibrator Circuits. *Electronic Engng.* *37*, 182 (1965).

Automatic Gain Control for Noise

By J. S. Bowman* and H. Sutcliffe*

The application of a.g.c. to noise amplifiers is complicated by the unsuitable characteristics of detectors with pre-detection delay. An analysis is given of the response of such detectors when the noise is normally distributed, and the conclusion is reached that post-detection amplification and delay is necessary for effective a.g.c. A circuit is described for control of a noise-generating amplifier in the 10 to 100kc/s region. The variable-gain element is a diode potential divider with controlled current bias.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

THERE is an increasing tendency towards the use of random noise as a test signal and this article discusses a problem encountered by the authors while working in this general field. A random noise source was required with a flat spectrum between 10kc/s and 100kc/s, providing a noise p.d. of r.m.s. amplitude close to 1V in a resistance 2kΩ. A circuit consisting of four transistors in cascade in the common emitter configuration, with associated *CR* networks, provided the basis for a suitable source. The first stage noise was amplified to a more than adequate level and a manual gain control could be adjusted to reduce the output level to the required 1V. The further refinement of adding automatic gain control (a.g.c.) proved to be not altogether straightforward, and forms the subject of this article.

Automatic control of a signal requires two separate processes. The first of these is detection, that is the production of a control signal, a current or p.d., dependent on the quantity to be controlled. The second process is the alteration of gain or attenuation in the main signal path, such alteration being effected by the control system. The overall degree or 'tightness' of control is particularly effective if the well-known principle of delay is exploited. The principle can be illustrated by one particular example. The quantity to be controlled is the r.m.s. value σ of the output p.d. of the amplifier. With no control the estimated value of σ would lie anywhere between 1V and 20V, because of the range of tolerances of the devices and components. σ is required to lie between 1 and 1·1V, so the control signal should have no effect for $\sigma \leqslant 1$V while for $\sigma = 1{\cdot}1$V the control signal must reduce the amplification by 20 times or 26dB. The delay in this context is 1V. The word 'delay' is here quite unrelated to time-delay, a more familiar usage, so the terminology is unfortunate.

The usual form of delayed a.g.c. circuit commonly used in radio receivers employs a biased diode detector. The cathode is held at a potential several volts positive with respect to mean anode potential, so that rectification and gain control only occur when the radio frequency signal exceeds the bias p.d. This type of delay will be called 'pre-detection delay', and it will be shown that this type of circuit is not very effective when noise level is to be controlled. This topic is discussed in the following section, while a later section describes a complete a.g.c. circuit that proved to give a satisfactory performance. The circuit uses 'post-detection delay' and incorporates a directly connected amplifier.

Detection of Normally Distributed Noise

A noise p.d. $v(t)$ produced from band-limited shot noise or resistance noise has a 'normal' or Gaussian distribution of amplitude. The probability of $v(t)$ lying between v and $v + dv$ is given by:

$$P(v)dv = (2\pi\sigma^2)^{-\frac{1}{2}} \exp(-v^2/2\sigma^2)\,dv \quad \ldots\ldots \quad (1)$$

where σ is the r.m.s. value of $v(t)$.

Consider now the behaviour of a simple rectifier circuit using pre-detection delay, as in Fig. 1. V_d is the delay voltage. The rectified or output p.d. is v_o and for a.g.c. purposes the mean value $v_{o(mn)}$ is of interest. Equation (1) shows that there is no limit to the range of values of $v(t)$, so there is no relation between σ and V_d that produces

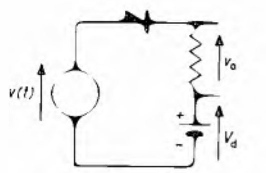


Fig. 1. Basic detector with pre-detection delay

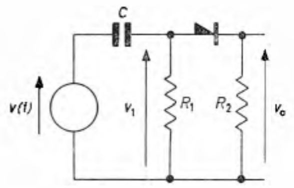


Fig. 2. Example of a detector circuit in which pre-detection delay occurs

zero value for $v_{o(mn)}$. A definite delay cannot be produced in this manner. An analysis of the circuit is still of interest however, because most practical rectifying circuits introduce a delay voltage. Two factors may contribute to the delay, the characteristic $I - V$ relation of the diode and the particular form of the circuit configuration. Fig. 2 is an example of a circuit in which delay is introduced. Inspection of the figure shows that, provided C is adequately large, the input p.d. to the rectifier circuit is given by:

$$v_1 = v(t) - v_{o(mn)}\,R_1/R_2 \quad \ldots\ldots\ldots\ldots \quad (2)$$

The last term in this equation is the voltage delay. It is of course not of constant value since it depends on the strength of the signal.

The above preamble shows that an understanding of the performance of a practical rectifier requires an analysis of the behaviour of the simple circuit of Fig. 1. Assuming the diode is perfect, then:

If $v(t) < V_d$, $v_o = 0$

If $v(t) > V_d$, $v_o = v(t) - V_d$

The waveform $v(t)$ occupies the range v to $v + dv$ for a proportion $P(v)\,dv$ of a long interval of time, so the mean value can be computed from:

$$v_{o(mn)} = \int_{V_d}^{\infty} (v - V_d)\,P(v)\,dv \quad \ldots\ldots\ldots \quad (3)$$

The normal probability distribution given by equation (1) is assumed, and the following expression can be derived for the mean output p.d.

$$v_{o(mn)} = \sigma\,(2\pi)^{-\frac{1}{2}}\,\{\exp(-y^2) - y\pi^{\frac{1}{2}}\,[1 - \mathrm{erf}(y)]\} \quad \ldots\ldots\ldots\ldots \quad (4)$$

where $y = V_d/\sigma\,2^{\frac{1}{2}}$, erf($y$) is a tabulated function[1].

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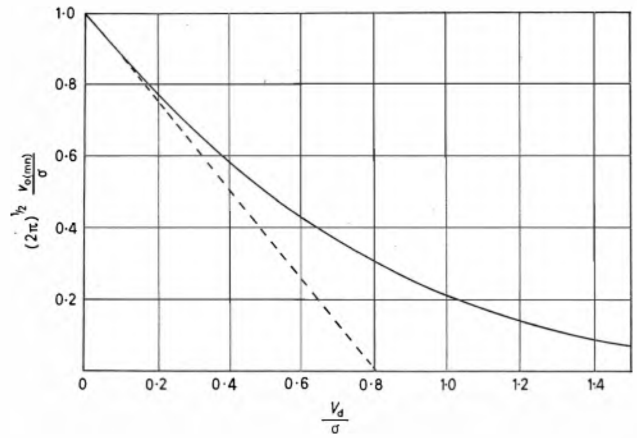


Fig. 3. Response of Fig. 1 as described by equation (4)

A graph of equation (4) is given in Fig. 3. The dashed line in the figure is the tangent to the curve as V_d approaches zero compared with σ, the slope being such that $dv_{o(mn)}/dV_d = -0{\cdot}5$ at $V_d = 0$.

When the delay voltage is negligibly small compared with σ, equation (4) reduces to the simple form:

$$v_{o(mn)} \to \sigma\,(2\pi)^{-\frac{1}{2}} \text{ as } y \to 0 \quad \ldots\ldots\ldots\ldots (5)$$

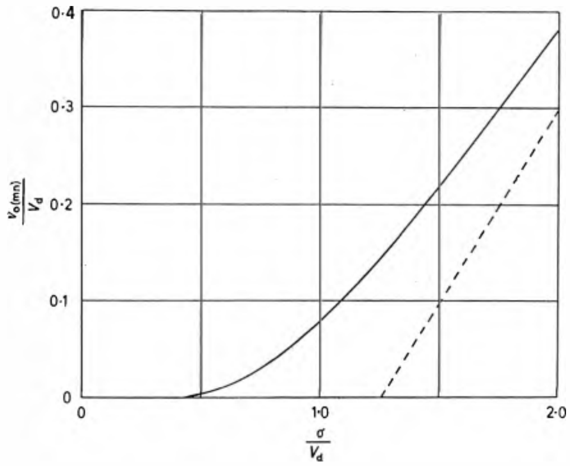
The rectified mean voltage is then linearly proportional to the r.m.s. noise p.d., a result with important general implications as explained in the Appendix.

Fig. 3 gives the relation between σ, V_d and $v_{o(mn)}$ in the form relating directly to equation (4). An alternative form of the relation is illustrated in Fig. 4, which shows $v_{o(mn)}/V_d$ as an explicit function of σ/V_d. The asymptote as σ/V_d becomes large is shown as a dashed line, and is given by:

$$v_{o(mn)}/V_d \to 1/(2\pi)^{\frac{1}{2}}\,[\sigma/V_d - (\pi/2)^{\frac{1}{2}}] \text{ as } \sigma/V_d \to \infty \quad \ldots\ldots\ldots\ldots (6)$$

It can be deduced from Fig. 4 that pre-detection delay does not produce an output-input characteristic of suitable form for effective a.g.c. The growth of $v_{o(mn)}$ as σ increases is gradual and lacks the necessary abruptness.

Fig. 4. Alternative form of displaying the response of Fig. 1



An A.G.C. Circuit with Post-Detector Delay

The particular circuit to be described in this section uses a pair of point-contact germanium diodes as a variable potential divider between the second and third stages of the four-stage amplifier. The incremental resistances of the diodes are determined by a delayed control current derived from a transistor functioning as a combined detector-amplifier. The circuit is illustrated in Fig. 5.

Diodes MR_1 and MR_2 are the principal components of the two arms of the potential divider. In normal operation both diodes are forward biased and under these conditions each has a current-voltage ($I - V$) characteristic with the approximate form:

$$I \simeq I_s \exp\,(V/V_o)$$

where I_s, V_o, are constants for the diode type. The parameter V_o is in the region of 25mV at room temperature. The incremental resistance is seen to be a sensitive function of current, being V_o/I. Inspection of the circuit shows that the sum of the currents in the two diodes is maintained by R_3 at approximately 50μA, so the impedances of the two diodes are affected by the controlling current I_2 as follows.

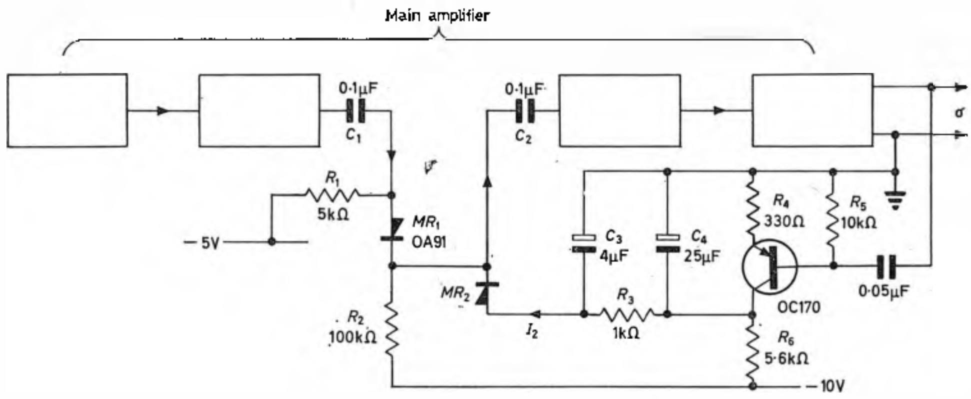


Fig. 5. A.G.C. circuit with post-detection amplification and delay

Let Z_1 and Z_2 be the incremental resistances of the diodes.

As I_2 goes from 0 to 50μA,

Z_1 goes from 500Ω to infinity,

Z_2 goes from infinity to 500Ω.

An exact analysis of the attenuating properties of the potential divider requires a detailed description of the diode characteristics and a discussion of the adjacent impedances in stages 2 and 3 of the main amplifier. The particular circuit used by the authors gave a total increase in attenuation of 40dB in the main signal path, associated with a change of potential of approximately 0·5V at the anode of MR_2.

The control current is supplied by the OC170 transistor, which functions as a detector and amplifier. The input noise p.d. of r.m.s. value σ is applied between base and ground, and rectification taken place at the base-emitter junction. The presence of R_4 in the emitter circuit gives a twofold advantage. It increases the input impedance and also ensures that the performance of the circuit is almost independent of transistor characteristics. The mean rectified current in R_4 is estimated as follows. Assume σ is 1V and that the base-emitter junction imposes a pre-detection delay of 0·2V. Then from Fig. 3 it can be deduced that the mean p.d. across R_4 is 0·3V. The current gain of the transistor from emitter to collector is close to unity, so the p.d. across the collector load R_6 is very nearly 5V. Under these circumstances current I_2 is finite, that is the a.g.c. circuit is within its operative region. The degree of tightness of control may now be assessed by considering the properties of the attenuator.

The attenuator swings through 40dB as I_2 increases from 0 to 50μA. To produce this current, the rectified current through the transistor must increase by 140μA, since the 0·5V change across the attenuator implies an increase of 90μA through R_6. Thus the rectified current increases by less than 16 per cent. The proportional increase in σ required to produce these changes is less than 16 per cent. It may be concluded that provided the uncontrolled value of σ lies within $\pm$20dB of nominal the controlled value will lie within $\pm$8 per cent or $\pm$0·7dB of nominal. This degree of tightness of control was considered to be adequate for the particular requirements of the authors. If it were necessary to establish an exact value of σ, a slight adjustment could be made to the value of R_4 or of R_6.

Discussion

Some explanation is needed to excuse the use of what appears to be a complicated circuit to perform a simple function. In particular it may be thought that a preferable technique would be to vary the gain of the amplifier by control of the bias of one or more of the transistor stages. The objection to this is that it would then be necessary to hold the relevant emitter potentials at some value intermediate between the supply line potentials, and so would call for either heavy current bleeds or reference diodes. The corresponding delay potential in Fig. 5 is at a point where the current change is only 50μA.

Another apparently preferable technique would be to dispense with the transistor in the a.g.c. loop and use a simple diode detector with post-detector delay. The difficulty here is that the delay voltage would still have to be in the region of 5V to achieve the same tightness of control. This would call for an output stage of the main amplifier capable of producing a value of σ in the region of 12·5V if a half-wave rectifier was used as the detector. Post-detection amplification of the a.g.c. signal is clearly preferable, and it is thought that the circuit of Fig. 5 is not so extravagant in the use of components as would appear at a first inspection.

Appendix

Measurement of r.m.s. Noise with a Common Type of Voltmeter

Many commercial electronic voltmeters incorporate an amplifier followed by a full-wave rectifier bridge, and use the current through the bridge as a feedback signal to stabilize the gain. Instruments of this type are free of rectifier delay and their response is therefore proportional to the mean magnitude of an input signal. When the input signal is normally distributed noise of r.m.s. value σ the response is proportional to σ. Instruments of this type are usually calibrated to present the r.m.s. value of sinusoidal input signals, so when the instrument reads normal noise the scale factor may be calculated as follows.

The instrument reading, V_r, is $\pi/2\sqrt{2}\ |v_{in(mn)}|$ where v_{in} is the fluctuating component of the input signal. When v_{in} is normal noise of r.m.s. value σ, $|v_{in(mn)}|$ has twice the value given by equation (5), the factor of two arising because there is a full-wave rectified circuit. The instrument reading V_r is given by:

$$V_r = (\pi/2\sqrt{2}) \cdot (2\sigma/\sqrt{(2\pi)}) = (\sqrt{\pi/2})\,\sigma$$

that is: $\sigma \simeq 1{\cdot}13\,V_r$

This result is not, perhaps, as generally known as its usefulness warrants. It is valid so long as the following precautions are taken.

(1) The detector is of a true 'mean magnitude' type.

(2) The instrument reading is not more than one-third of full scale, to guard against saturation at noise peaks in the preamplifier.

(3) The bandwidth of the instrument is adequate for the frequency spectrum of the noise under investigation.

(4) The appropriate scale factor is applied. The factor is 1·13 for normally distributed noise.

There is another common type of electronic voltmeter that uses an envelope or peak detector followed by a d.c. amplifier. Instruments of this type are not convenient for noise measurement because the response depends on the value of CR time-constants associated with the detector.

Reference

1. Flugge, W. Tables of Transcendental Functions, p. 110. (Pergamon. London, 1954).

A Pencil Follower Trace Analyser

The pencil follower tracer analyser manufactured by d-mac Ltd, Queen Elizabeth Avenue, Glasgow, S.W.2, is intended for analysing curves and providing XY co-ordinates in a form suitable for input to a digital computer.

The trace, which may be in the form of a paper chart or film or maps, engineering drawings, etc., is placed on the hard white plastic top, or if on photographic film, projected by an optical system on to the table-top. The material is kept flat by a thin, transparent, holding down sheet.

The user employs a 'pencil.' A light wire connects the 'pencil' to a source that energizes the 'point' of the pencil at 400c/s and creates there an a.c. magnetic field. A servo mechanism, or follower under the table-top tracks the 'pencil' wherever it may move over the surface of the table. The follower is attached to X and Y electromechanical digitizers. The latter are, in turn, connected to X and Y encoders in an auxiliary console.

The servos and digitizer units are housed for easy servicing under the table top.

X and Y digital displays are sited on the console, which may also house the fixed address, counter, timer for line work, and output serializer units. The latter feeds into the interface unit through to the desired output device.

A free-moving keyboard attached to the console is used to instruct the system. The keyboard may be placed anywhere on the table at the convenience of the user.

Thus the user simply points his 'pencil' for point information on curves, shapes, diagrams and so forth, or traces the curve if he requires continuous line information, and presses a button or footswitch to operate the system and obtain a reading. There are several types of 'pencil' for different uses.

The table has a working area of 18 by 40in. The accuracy of the type PF 1000 is 1·0mm per digit and the PF10 000, 0·1mm per digit. When continuous records are read, the maximum width of roll paper is 18in and motorized or manually operated paper reels may be added to each end of the table to facilitate paper or continuous record material handling.

The projected images may be obtained from 16 or 35mm continuous film or from 24 by 24 and 36 by 24mm frames, in lengths up to 300m.

The output interface depends upon the type of output device, such as card punches, tape punches and typewriters. The serialization of output is very flexible with two ten-position output rings A and B, and a third, C, acting as control for the selection of ring A or B. Each ring may be plugged as required by the operator, and control characters may be added to obtain the desired format. Special interfaces (and encoders where required) are available for many output devices.

A Nitrogen-cooled Degenerate Parametric Amplifier for a Radiometric Application

By D. Chakraborty*, B.Sc. and G. F. D. Millward*, B.Sc. (Eng.)

The design of a 6Gc/s liquid-nitrogen cooled degenerate parametric amplifier for use in a radiometer for communication-satellite earth-station aerial gain measurements via a radio star is described. Two stages, each of 15dB gain and 200Mc/s bandwidth to the −3dB points, are cascaded to provide a minimum gain of 30dB. The inherent bandwidth of the amplifier was adequate for the purpose and special circuits to broaden the bandwidth were not required. The noise temperature of the cooled amplifier is 57°K.

The theoretical limit of the sensitivity of a radiometer system using the amplifier is briefly discussed.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

IN a communication-satellite system knowledge of the earth-station aerial gain achieved in practice at both the transmitting and receive frequencies is essential. With large aerials the measurement of aerial gain using sources near the ground is subject to errors and an independent method of measurement is desirable. One of the most useful methods is to evaluate the aerial gain by detecting the noise power from a radio star whose power flux density is known. At the receive frequency, e.g. 4Gc/s, such measurements can be made via the normal low-noise receiving system, but at the transmit frequency, e.g. 6Gc/s, an additional low-noise amplifier is required. For the latter purpose a liquid-nitrogen cooled degenerate parametric amplifier is a convenient amplifier.

In a degenerate parametric amplifier the input circuit is tuned to half the pump frequency f_p so that the signal and the idler frequency channels coincide. If the input signal consists of wide-band noise, then for a symmetrical amplifier characteristic with the centre of the pass-band coinciding with $\frac{1}{2}f_p$, the amplified noise signal below $\frac{1}{2}f_p$ will have complete correlation with the amplified noise signal above $\frac{1}{2}f_p$. This correlation, which is due to the complementary nature of the signal and the idler frequency transformations, leads to an increase in the post-detector fluctuations. When the incoming signal is in the form of noise the degenerate parametric amplifier offers the logical choice for the best sensitivity. It has been shown[1] that the double-sideband noise temperature of a degenerate parametric amplifier is almost half the noise temperature that would be available from a non-degenerate parametric amplifier having the same diode junction properties and operating bias. Nevertheless, with a diode at ambient temperature, the calculated amplifier noise temperature was too high. The amplifier was therefore designed for operation at 77°K. Gallium arsenide varactor diodes suitable for operation at this temperature are commercially available.

To reduce the noise contribution from the mixer and the following stages of the system, a minimum r.f. gain of 30dB is required. To increase stability this is best provided by two 15dB stages in cascade. As the application does not call for a large bandwidth no special circuits were needed to widen the 3dB bandwidth beyond 200Mc/s, this being readily obtained from each 15dB stage.

The amplifier is basically a one-port device and an additional external component is required to separate the input and output signals. A ferrite circulator is usually used for this purpose, providing adequate isolation between input and output, with only a small additional attenuation. The attenuation in the circulator will contribute a further component to the noise level of the system, and under ideal conditions, it is desirable for the circulator to be closely attached to, and cooled with the amplifier. Unfortunately 6Gc/s circulators suitable for operation at 77°K are not at present available, and it was necessary to use uncooled circulators. The connexion between the amplifiers and their circulators is provided by sections of thin-walled stainless steel waveguide, about 2ft in length, to minimize heat transfer. However, the inclusion of this length of waveguide in the signal circuit reduces the bandwidth and increases the noise temperature considerably, when compared with values calculated from the simple theory of parametric amplifiers.

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Theoretical Considerations

The basic theory of degenerate parametric amplifiers is described in several recent publications[1,2] and it is only necessary therefore to mention the relations needed for design calculations.

For a reflection type of amplifier using a circulator, the power gain[1] g is given by:

SYMBOLS

β	=	phase constant = $2\pi/\lambda$
γ	=	coefficient of capacitance modulation
f_c	=	cut-off frequency of the varactor diode
f_p	=	pump frequency
f_s	=	signal frequency
g	=	power gain of the first amplifier
P_p	=	pump power
Q	=	Q-factor of the diode = f_c/f_s
R	=	magnitude of the negative resistance as seen by the passive signal circuit
R_g	=	generator resistance
R_s	=	spreading resistance of the varactor diode junction
R_T	=	total signal circuit resistance
T_{dsb}	=	double-sideband noise temperature
T_d	=	physical temperature of the diode
T_2	=	double-sideband noise temperature of the second amplifier
Z_o	=	characteristic impedance of an arbitrary transmission line
Z_d'	=	normalized driving point impedance
Z_1	=	diode terminal impedance

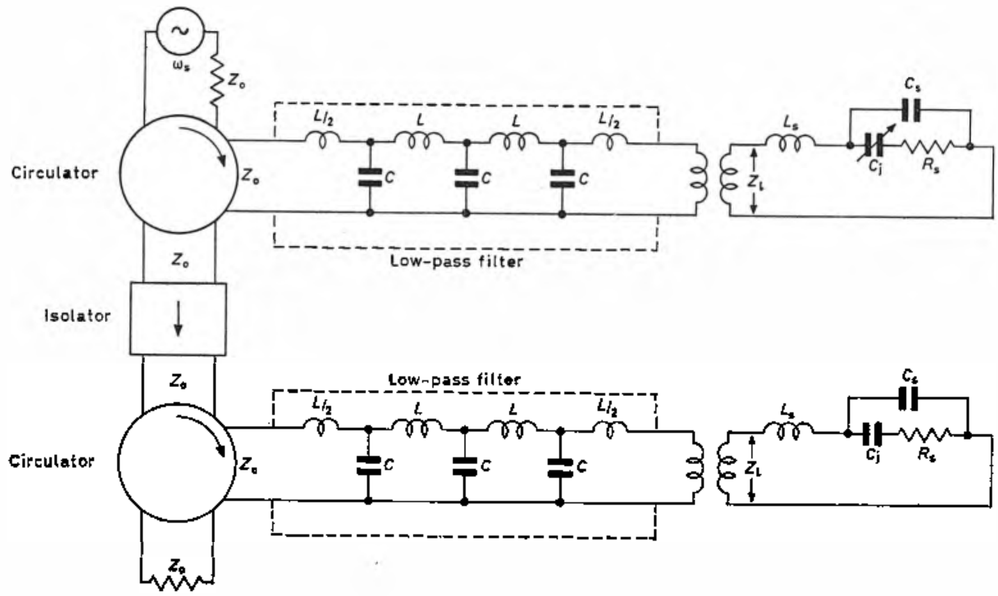


Fig. 1. Circuit representation of two cascaded negative-resistance degenerate parametric amplifiers

$$g = \frac{4R_g^2}{(R_T - R)^2} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

R_T represents the signal circuit resistive loss, R_g the generator resistance as seen by the diode.

$$R = F\,[(\gamma Q), P_p] \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

R is the magnitude of the negative resistance as seen by the passive signal circuit under the pumped condition. R is only partially amenable to analysis, but an exact knowledge of it is not essential for the design of the parametric amplifier. Equation (2) shows the functional relation between R and the dynamic quality factor of the diode (γQ), and pump power.

The circuit representation of two cascaded negative resistance degenerate parametric amplifiers is shown in Fig. 1. The perturbation in the signal circuit impedance caused by the pump circuit coupling is neglected. With a small signal the non-linear element of the junction capacitance can be represented as a time varying linear element and expressed as the first two terms of a Fourier Series:

$$C(t) = C_o\,(1 + 2\gamma \cos \omega_p t) \quad \ldots\ldots\ldots \quad (3)$$

Here the contributions due to higher order sidebands and harmonics are considered to be of no significance. The term C_o represents the fixed reactance due to the stray elements and static junction capacitance (L_s, C_s and $C_J(0)$, in Fig. 1), which stores a considerable amount of signal energy and limits the bandwidth capability of the amplifier. Various techniques[2,3] have been proposed to improve the bandwidth capability by using bandpass filter circuits with appropriate reactance slopes and utilizing the static reactance of the diode as part of the filter circuit. Although this technique would have improved the bandwidth capability of the amplifier considerably, this was not required for the present application.

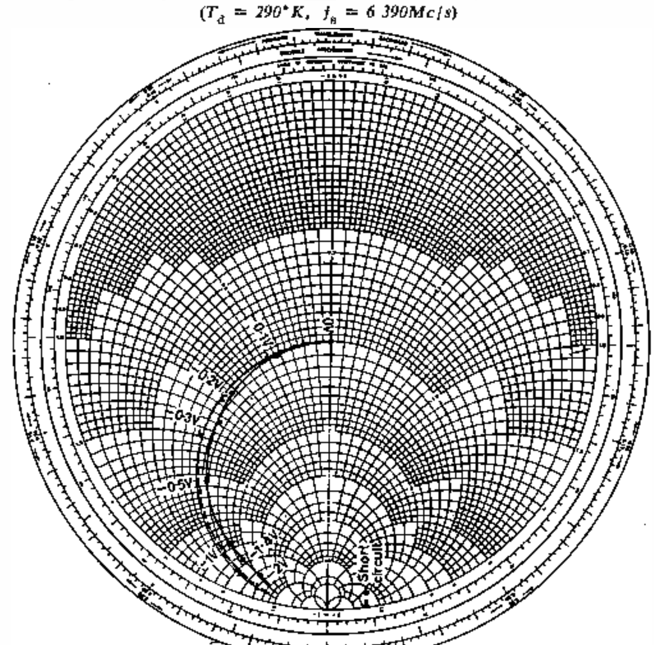
Practical Design of the Amplifier

The following simple method has been used in the design of this amplifier. Referring to Fig. 1, a low-pass filter in 50Ω coaxial line is incorporated in the signal circuit in order to stop pump leakage to the second stage and the proper generator resistance is provided by a linearly tapered transformer. The value of R_g determines the diode loading and hence the effective noise temperature, the optimum value of R_g for minimum noise temperature being given by:

$$R_g = (\gamma Q - 1)\,R_s \quad \ldots \quad (4)$$

The Q of the diode is determined experimentally by mounting the diode in a coaxial line. The diode is matched at zero voltage bias using an impedance transformer. The reverse bias voltage is then gradually changed and the points as shown in Fig. 2 are plotted from the measurement of phase and the magnitude of the resulting standing-wave pattern. A short-circuit point is established by replacing the diode with a dummy. The impedance locus thus obtained lies inside the unit-resistance circle. This, coupled with the fact that the short-circuit point lies away from the periphery of the Smith's chart, indicates the presence of distributed loss in the diode circuit. Had the diode been mounted in a hypothetical circuit with a lossless transformer the impedance locus would have been superimposed upon the unit-resistance circle and the short-circuit point would lie on the periphery of the chart. Hyde and Smith[4] have discussed the effect of loss in degrading the diode Q measurement. However, the loss of the circuit including the matching transformer can be experimentally determined and the necessary corrections are introduced to bring the experimental points closer to the unit-resistance circle. The Q-factor is the ratio of the imaginary part to the real part of the diode impedance. The Q at a particular bias is therefore given by the mag-

Fig. 2. The impedance locus for the diode at various reverse biases
—●— *Experimental;* —▲— *Corrected for loss conductance*
(T_d = 290°K, f_s = 6 390Mc/s)



nitude of the reactance swing from the short-circuit point to the corresponding bias point, as the real part lies on the unit circle. In Fig. 2 it will be observed that the zero-bias Q for the diffused-junction gallium arsenide varactor used is 13·5.

The spreading resistance of the diode is 3Ω and an estimate of the value of γ is 0·3. From equation (4) the computed value of R_g is thus approximately 9Ω.

A detailed diagram of the amplifier is shown in Fig. 3. The diode is held in the reduced-height pump waveguide by spring loading, and the mount holding the diode is given a thin p.t.f.e. coating to ensure d.c. insulation. The pump power is matched to the diode by a stepped transformer. A single klystron is used as a common pump source to both amplifiers through a Y-splitter and isolators. The klystron is maintained in an oil bath for stability.

The complex impedance Z_1 at the diode terminals can be transformed by an appropriate electrical length l between the diode terminals and an arbitrary reference plane so that the transmission line equation:

$$Z_d' = \frac{Z_1 + jZ_o \tan \beta l}{Z_o + jZ_1 \tan \beta l} \quad \ldots\ldots\ldots\ldots (5)$$

yields a real solution for the normalized driving point impedance Z_d', where Z_o is the characteristic impedance of the signal-circuit waveguide. This enables the signal circuit to see an almost resistive impedance at the signal frequency band. The three-screw tuner in the signal-circuit waveguide provides additional tuning and enables the diode to be tuned to resonance when cooled to liquid-nitrogen temperature. The gain frequency response of a single amplifier is shown in Fig. 4.

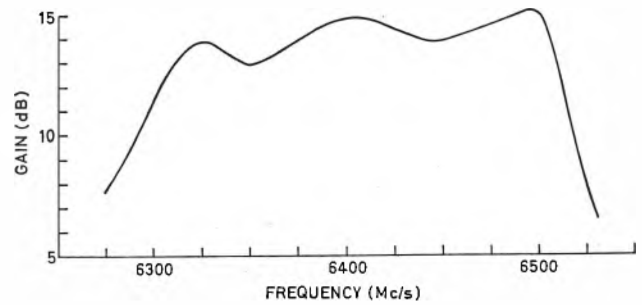


Fig. 4. Gain of a single amplifier

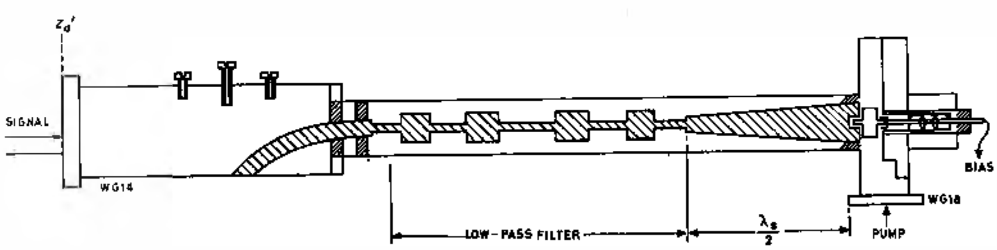


Fig. 3. A 6Gc/s degenerate parametric amplifier

Cryogenics

The two separate amplifiers are mounted together and enclosed in a copper can and this can is mounted inside a double-walled metal Dewar which can be filled with liquid nitrogen, boiling at 77°K. The complete assembly of the amplifier is shown in Fig. 5. A four-port waveguide switch is incorporated in the input of the amplifier for noise measurement and the two symmetrical bends in waveguide 14 comprise the signal circuits for the amplifiers. The ambient temperature circulators assembly is enclosed in the hexagonal aluminium box.

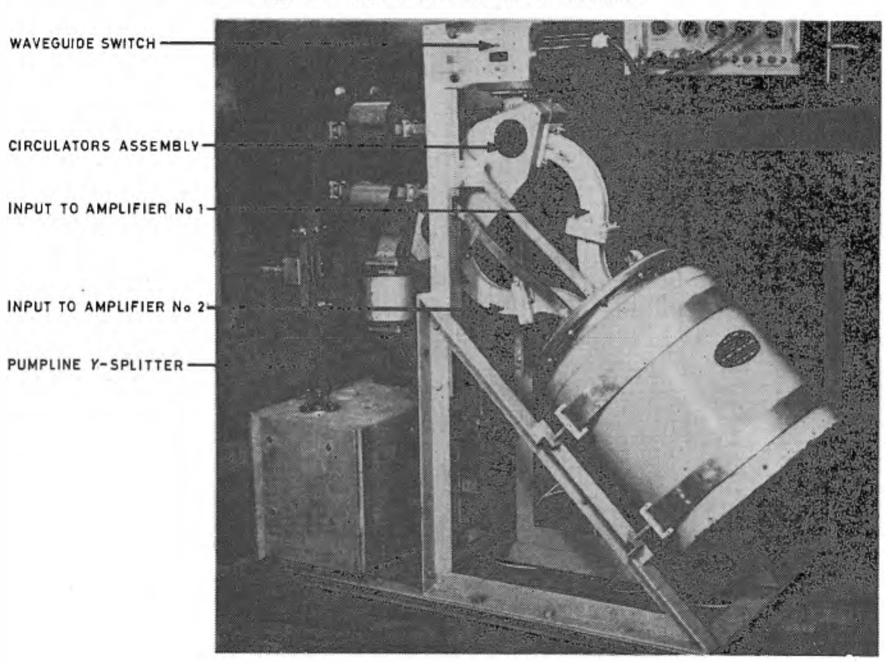
The pump klystron is enclosed in an oil bath (not visible) inside the aluminium box shown on the left. The output from the cascaded amplifier is fed to the radiometer unit (not shown) via a waveguide circulator as shown in a cylindrical aluminium box. The double-walled metal Dewar is strapped to the main structure at an angle so that the inclination does not exceed more than 45° for any angle of elevation of the aerial between the horizon and the zenith.

A dry-nitrogen pressure transfer technique is used for transferring liquid nitrogen into the Dewar at regular intervals. In order to prevent leakage of refrigerant, the inside of the copper can is pressurized with gaseous helium at some 2lb/in². Low loss microwave windows are provided in the signal and the pump line waveguides to prevent this helium escaping.

Noise Considerations

A degenerate parametric amplifier possesses two noise temperatures, namely the single-sideband noise temperature and the double-sideband noise temperature. If the incoming signal is in the form of noise and the amplifier passband is centred on $\frac{1}{2}f_p$, then the noise amplitude functions below $\frac{1}{2}f_p$ and above $\frac{1}{2}f_p$ are independent slowly fluctuating quantities having equal r.m.s. values and equal normalized auto-correlation functions. Under these condi-

Fig. 5. Complete assembly of the amplifier



tions the d.s.b. noise temperature is the lowest and the noise temperature equation for the amplifier can be written in terms of the diode parameters:

$$T_{dsb} = \frac{T_d}{(\gamma Q - 1)} + (T_2/g) \qquad (6)$$

where T_2 is the noise temperature of the second amplifier, and g is the gain of the first amplifier. Equation (6) does not take into consideration the noise contributed from the circulator and the finite loss of the signal circuit. The ambient temperature circulator contributes some 14°K and the stainless steel waveguide which is part of the signal circuit and the microwave switch in front of the first amplifier together contribute some 10°K. With the diode at liquid-nitrogen temperature (77°K) the theoretical d.s.b. noise temperature should be about 51°K. The noise temperature of the amplifier was experimentally determined by hot and cold load technique (the hot termination at 290°K and the cold termination at 77°K). The measured value was 57 ± 5°K.

Radiometric Application

A typical noise-adding radiometer comprises a pre-amplifier (with provision for the injection of some additional noise from a noise tube through an appropriate directional coupler) followed by an i.f. amplifier, high-level square-law detector, ratiometer, low-pass *RC* filter and display unit. From the theory of a square-law detector and the theory of operation of the ratiometer it can be shown[6] that the ultimate sensitivity of the radiometer is:

$$\Delta T_s/T_s = \frac{(T_s + T_A)}{T_A} \pi/2 \;\; 1/\sqrt{(B\tau)} \qquad (7)$$

where

T_s is the system noise temperature (i.e. feeder loss temperature plus the amplifier noise temperature),

T_A is the additional noise from the noise lamp,

B is the i.f. bandwidth (square shape),

τ is the post-detection filter time-constant.

For a system having $T_s = T_A = 100$°K,

$$B = 25 \times 10^6 \text{c/s},$$
$$\tau = 1\text{sec}$$
$$\Delta T_s = 0{\cdot}0628\text{°K} \qquad (8)$$

From equation (7) it will be observed that any drift in the pre-amplifier gain will produce a corresponding drift in T_s and ΔT_s will become uncertain. This is eliminated by pulsing the noise tube on and off and providing appropriate balancing to restore the fluctuations due to incoming noise. The second uncertainty is the variation of noise-tube output which cannot be eliminated. The variation in noise tube output degrades the theoretical sensitivity to some extent but 0·1°K may be considered as a realistic practical limit.

Conclusion

A low-noise degenerate parametric amplifier, suitable for use in a noise-adding radiometer for the measurement of the gain of earth-station aerials, using Cassiopeia-A as the external reference source, has been described.

Acknowledgments

The authors wish to thank their colleagues, both in the Post Office Research Station and at Goonhilly Radio Station, for their co-operation. The permission of the Engineer-in-Chief of the Post Office to make use of information contained in this article is gratefully acknowledged.

REFERENCES

1. Blackwell, L. A., Kotzebue, K. L. Semiconductor-Diode Parametric Amplifier, p. 58 and pp. 67-72. (Prentice Hall Inc., London, 1961).
2. Siedel, H., Herrman, G. F. Circuit Aspect of Parametric Amplifier. 1959, I.R.E. Wescon Convention Record. Pt. 2, pp. 83-90.
3. Matthaei, G. L. A Study of the Optimum Design of Wideband Parametric Amplifiers and Upconverters. Trans. I.R.E., M.T.T. (January 1961).
4. Hyde, F. J., Smith, R. B. Effect of Losses in Varactor-diode Impedance Measurement. *Proc. Inst. Elect. Engrs. 111*, 471 (1964).
5. V.H.F. Techniques, Radio Research Laboratory, Harvard University. (McGraw Hill Book Co. Inc., 1947).
6. Ohm, E. A., Snell, W. W. A Radiometer for a Space Communication Receiver. *Bell Syst. Tech. J. 42*, 2047 (1963).

Electronic Data Display for Air Traffic Control

One of Britain's most advanced experiments in air traffic control, at Prestwick in Scotland, is to have five Marconi electronic data displays which can 'write' characters and symbols on the face of a cathode-ray tube at the rate of 50 000 characters per second. This type of display will provide the fastest and most flexible flight data presentation system in the world.

The Prestwick experiments are based on the use of a Ferranti APOLLO computer to process information on aircraft movements, in order to present the appropriate data to air traffic controllers more rapidly and more effectively than is otherwise possible. This type of automation in air traffic control will help controllers to forecast conflictions more quickly and will give them the capability of controlling the greatly increased air traffic which is expected by the early 1970's.

The APOLLO computer project at Prestwick is being run by the National Air Traffic Control Services of the Ministry of Aviation and Ministry of Defence, to study the possibilities of using computer automated systems to handle data on the steadily increasing volume of air traffic. This particular project is principally concerned with the Oceanic Sector of the North Atlantic, which is controlled from Prestwick, but many of the results are expected to have a direct bearing on future air traffic control systems of all types.

The air routes over the North Atlantic are already heavily loaded at peak periods, with as many as 20 to 25 aircraft entering the sector per hour. This loading is certain to increase in the future, and the prime object of the APOLLO project is to investigate the extent to which the air traffic control functions can be automated by the use of a computer. Information is derived from a variety of sources; from aircraft already in the air, from flight plans transmitted by departure airports and air traffic control centres, and from the meteorological services. The computer can correlate this information rapidly and present controllers with the present position, and also with certain future predictions.

The Marconi displays will make it possible to present information in alpha-numeric form, as soon as it is available, to all the appropriate controllers. They will also make it possible to up-date this information instantaneously, and without the necessity for reprinting whole blocks of data.

Each character is written on the display tube face in 20μsec. A series of small deflexion voltages trace out the complete symbol on the tube face in response to 'writing' waveforms which are supplied by a character generator based on computer techniques.

By using this type of display, the same information can be made available, on separate tubes, to any number of different controllers at the same time. In the same way, parts of the data which relate to particular sectors of the overall airspace can be presented in an individually convenient form.

Of the five displays on order, four will have 8½in diameter display tubes, while the fifth will have a 17in tube.

In a typical air traffic control display, information can be arranged in tabular form, with each horizontal line of the table corresponding to a single aircraft in the control area. The columns of the table would then correspond to aircraft identity, and data such as destination and height, together with the estimated times of arrival at various reporting points and at the destination. During the progress of each flight, the actual flight times between the various points on the route will inevitably differ from the predicted flight plan. Up-to-date information derived from the aircraft and from radar can then make it possible to revise flight plans to fit the latest known facts.

Sampling System for High Degree Stabilization of D.C. Motors

By F. Szlávik* and Gy. Orbán*

A system has been developed for fully automatic start-up and high degree stabilization of speed of low and medium power d.c. motors. The electronic circuits described have been designed for a motor driving a neutron-beam chopper. The accuracy and stability in frequency and phase is determined, over wide ranges of the influencing parameters, by that of a reference signal generator.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

HIGH intensity monochromatic neutron beams required in experimental neutron physics are usually obtained by means of rotating mechanical selectors, so-called choppers, with curved slits. The neutron beam of wide energy spectrum coming from the reactor is chopped into neutron bursts of limited energy bandwidths, the centres of which are determined partly by the speed of the motor driving which the stability of the motor speed depends within wide ranges of values of the influencing parameters. Since the stability required under the given conditions was 0·3 per cent, a high precision l.f. signal generator (Muirhead D650-B) was thought to be adequate. For still higher accuracy one may use a quartz-controlled device necessitating a modification of the shaper stage only.

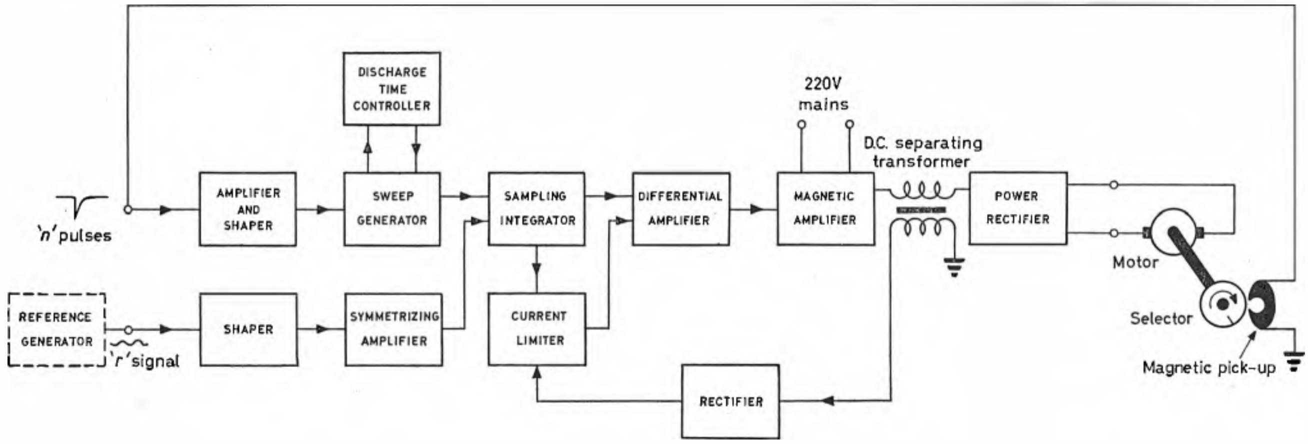


Fig. 1. Arrangement of control system

the selector. Since the overall accuracy of an experiment is markedly affected by the stability of the centre of the transmitted energy band, various systems have been developed to keep the speed of the driving motor constant. Merely stabilizing the speed averaged over a long period is not sufficient; it is the angular velocity that should be kept constant.

The stability aimed at by the present system was 0·3 per cent for motor speeds ranging from 2 000 to 5 000rev/min.

In most of the speed stabilizers used for the above purpose either classic phase discriminators, or digital control techniques are applied, the former requiring manual intervention in the process of synchronization, while the latter, as usual with digital devices, is rather complicated and costly. The relatively simple control system to be described is a combination of analogue and digital methods by which the motor can be brought from zero to the desired speed and kept running at this speed to a high degree of accuracy.

General Principles

The block diagram of the control system is shown in Fig. 1.

An external reference generator is used to deliver a signal of frequency f_r determining the desired speed of the motor. It is the frequency stability of this oscillator on

Pulses of frequency f_n proportional to the speed of the selector coupled to the driving motor are taken from a magnetic pick-up which delivers a voltage pulse induced in its coil at each turn of the selector by the 0·3mm thick magnetic plate fastened into the 200mm diameter jacket of the rotor. These pulses are of a width corresponding to 1/130 parts of the time taken by a complete turn of the rotor, thus they can be used also for phase stabilization.

The frequencies f_r and f_n are compared in a so-called sampling integrator stage to be described later. The difference appears as a d.c. error voltage at the input of the differential amplifier. Since the power consumption of the motor is rather high—about 300W for $n = 5\,000$rev/min—a magnetic amplifier has been incorporated. The a.c. supply is fed to the motor terminals through solid state power rectifiers.

The magnitude of the current fed to the motor is limited in order to increase motor life and to prevent possible runaway. To this end a voltage proportional to the motor current induced in the secondary of the d.c. separating transformer is passed through the current limiting amplifier to one of the differential amplifier inputs.

The Sampling System

The pulses n having their amplitudes and widths determined by the motor speed are first amplified by V_{1a} (Fig. 2) then converted by the Schmitt trigger circuit V_2 into pulses of constant amplitude. These pulses charge,

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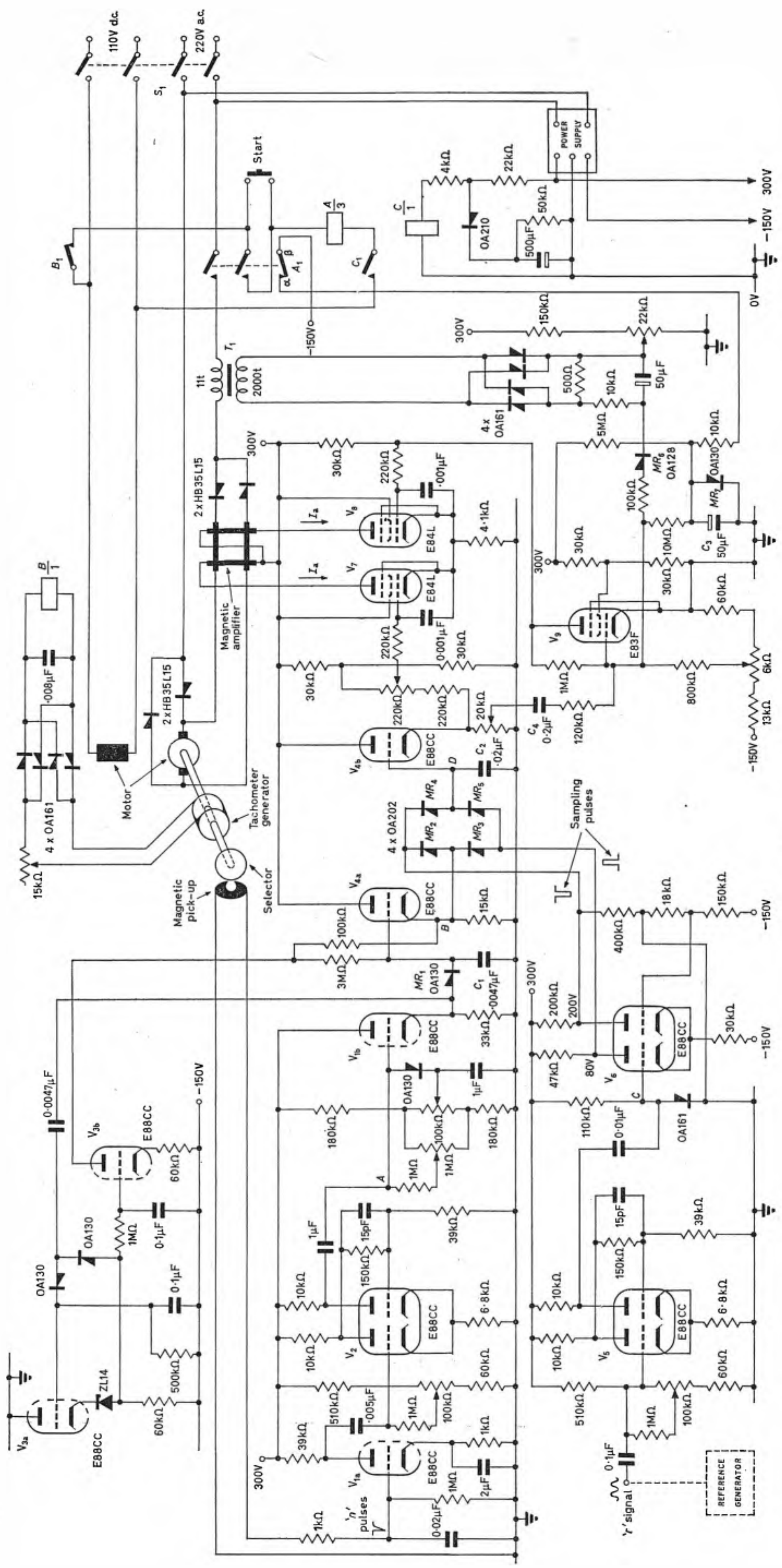


Fig. 2. Complete circuit of control system

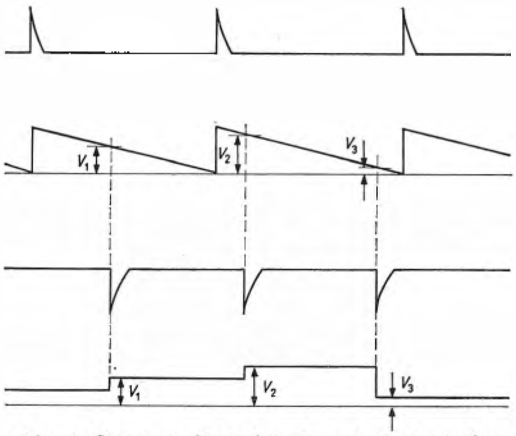


Fig. 3. Waveforms illustrating the method of sampling

through cathode-follower V_{1b} and diode MR_1, the capacitor C_1. The space factor of the sawtooth signal on this capacitor is kept at a nearly constant value by the discharge time controller circuit $V_{3a}V_{3b}$ which discharges the capacitor C_1 always to about 0V by the time the next charging pulse arrives, irrespective of the actual frequency of the n pulses. (Owing to the integrating circuit, the discharging current delivered to capacitor C_1 from V_{3b} is proportional to the instantaneous frequency of the pulses n.)

V_5, another Schmitt trigger circuit, squares the signal r from the reference generator which is passed after some further differentiation in the form of pulses with constant amplitudes and widths to the symmetrizing amplifier V_6. The diodes MR_2 to MR_5 when at rest, are backward biased by the anodes of V_6. However, on the appearance of a sampling pulse at the anodes, the diodes MR_2 to MR_5 of the bridge circuit conduct and charge the capacitor C_2 to the cathode potential of V_{4a}, that is, to that of the sawtooth generator capacitor C_1. The waveforms are shown in Fig. 3.

Between two samplings there is no change in the voltage across C_2 because of the high reverse resistance of diodes MR_2 to MR_5. This voltage is used to control, across the cathode-follower V_{4b}, the differential amplifier V_7V_8 and thus the motor current too. Since the possible values of the error signal determined by the amplitude of the sawtooth voltage are sufficient to cover the complete control voltage range of V_7V_8, no manual setting of the d.c. operating point for the desired speed is necessary.

Automatic Start-up

The above considerations seem to imply only the phase stabilizing potentialities of the system, i.e. torque correction corresponding to the difference between the instantaneous and the equilibrium phases, being possible if the frequencies of n and r are about equal. For frequency stabilization in a more general sense means are provided for by the limiter action of the differential amplifier V_7V_8, that is by the fact that past a given value of the control grid potential of V_7, its anode current cannot be further increased, since the cut-off of V_8 causes the total current from the cathode resistance, which is practically independent of the grid potential, to flow through V_7. This can be exploited to bring the motor, by a single action of a push-button, from zero to the synchronous speed and to maintain this speed essentially stable until switching off. Since the effect is equivalent to limiting the output voltage of the sawtooth generator, for simplicity, the procedure will be described as if that were the case.

Owing to the voltage limiting action, the sawtooth voltage becomes trapezoidal with the horizontal part about half as long as the instantaneous period T (Fig. 4). If the frequency of the pulses n proportional to the motor speed and that of the reference pulses r have no common integer multiple, any value of the trapezoidal voltage may be taken as sample. This means that, for example, for the idealized waveform in Fig. 4 averaging a few times over the period T, the controlling voltage of V_7, to which in a first approximation the motor current is proportional, gives:

$$V_{contr:av)} \simeq 3/4\ V_{max} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

(For symbols see Fig. 4). If this value is greater than that required to balance the loading torque at the instantaneous motor speed, the speed continues to increase.

It is easy to see that a sub-harmonic state becomes the most critical if:

$$f_n = f_r/2$$

which may cause a phase constellation (Fig. 4(a)) with

$$V_{contr} \simeq V_{max}/2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

To avoid inconvenient synchronization also in this case, one must have

$$V_{max} > 2\ V_{sy} \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

in the frequency range from $f_{r\,(min)}/2$ to $f_{r\,(max)}/2$ where $f_{r\,(min)}$ and $f_{r\,(max)}$ are specified minimum and maximum values of the reference frequency. Referring to Fig. 4(b), V_{sy} is the value of the control voltage required to keep the speed at a synchronous steady value.

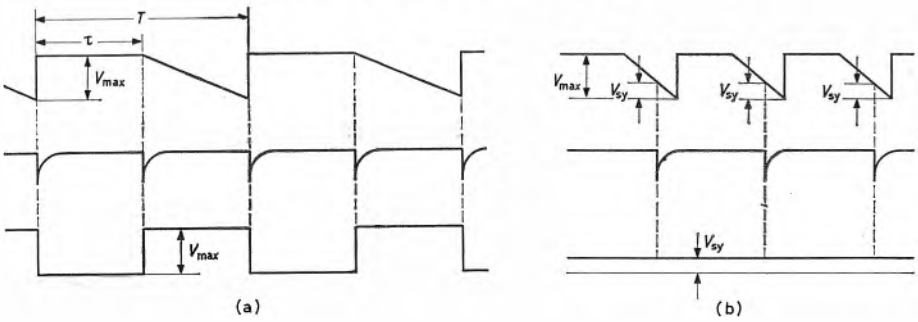
The condition required by equation (1) that

$$V_{max} > 4/3\ V_{sy} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

has to be satisfied from $f_{r\,(min)}$ to $f_{r\,(max)}$. One must not choose V_{sy} too low in order to permit the stabilizing effect to assert itself in the case of competing effects (e.g. mains fluctuation). It has to be emphasized that the above simplified considerations concerning the voltages should be applied by providing the necessary excess power for the synchronization of the motor. Therefore, the characteristics of both the magnetic amplifier and the motor have to be taken into account over the whole specified range of motor speed.

The system would work even if the ratio T/τ were chosen to be different from $T/\tau \sim 2$. Higher values, however, would require by equations (3) and (4) an over-dimensioning of the output stages, while lower values would result in less reliable synchronization.

Fig. 4. Sampling of limited sawtooth voltage



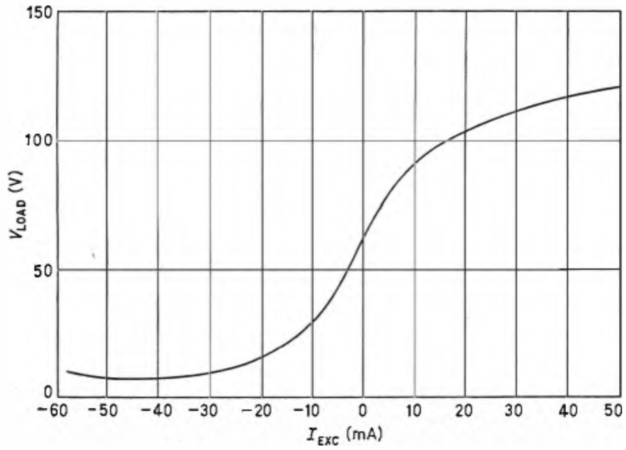


Fig. 5. Characteristic of the magnetic amplifier

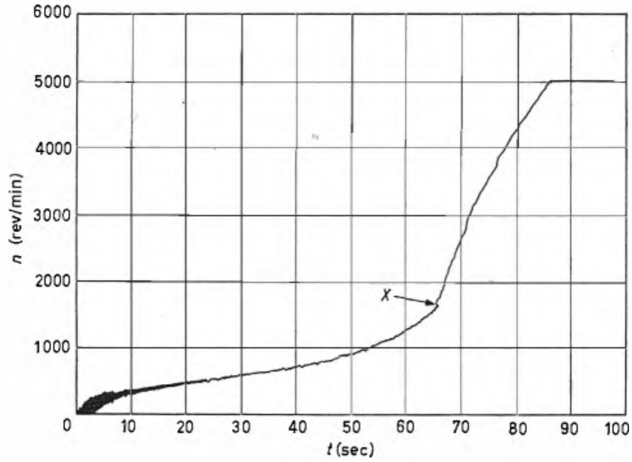


Fig. 6. Selector speed against the time elapsed since acting on the 'start' push-button (n_{sy} = 5 000 rev/min)

Further Circuit Characteristics

CORRECTION OF THE TRANSFER CHARACTERISTICS

The control system is of proportional type. Since the time-constant of the rotor is of the order of a few seconds and would induce a considerable time lag in control, a transfer characteristic correcting capacitor, C_4, has been added to render the control system a proportional-differential type. The value of C_4 is chosen to keep the transfer function just short of having singularities which might cause hunting of the selector.

SAFETY

Large current surges at the beginning and during the speed-up, likely to reduce motor life, are prevented by the contacts of relay *A* operated by the 'start' push-button. The contacts at rest are closed keeping the upper end of capacitor C_3 at about −150V, thus V_9 is partly cut-off. Consequently the control grid of V_8 is positively biased sufficient to block the other valve V_7 of the differential amplifier. The magnetic amplifier cannot be excited except by the anode current of V_8 thus, owing to the common characteristics of the power amplifier stages and the motor, the maximum possible value of the starting current is limited to 3A. The characteristic curve of the magnetic amplifier for a given ohmic load is shown in Fig. 5. The exciting current is: $J_{exc} = J_{aV7} - J_{aV8}$.

Operation of the 'start' push-button disconnects the contact pair α-β and causes capacitor C_3 to discharge thus permitting the error voltage (referred to above as control voltage) at the grid of V_7 to become more and more effective. V_9 will be fully conducting in about 90sec, i.e. in any case after the synchronization of the motor speed. The grid potential of V_8 is thereby sufficiently reduced to permit the total error voltage appearing at the grid of V_7 to be apparent in the currents of V_7—V_8 and to produce a strong synchronism maintaining effect.

Further current surges are prevented by a d.c. limiting voltage obtained by rectifying the output current pulses of the magnetic amplifier which are proportional to the motor current. This limiting voltage is d.c. separated by transformer T_1 and is passed on by diode MR_6 to the control grid of V_9 and thence to that of V_8, only if the motor current is greater than about 6A.

The motor is operated close to a nuclear reactor and a runaway due to circuit failure might cause serious damage. The quiescent mass of the rotating system is 11kg. Since in the case of shunt wound d.c. motors the interruption of the exciting (stator) current is the most likely to be responsible for a runaway, the current in the rotor is immediately disconnected by relay *A* if the exciting voltage decreases below a given value. In addition, an a.c. tachogenerator has been incorporated into the motor the voltage of which is proportional to the speed of the latter and acts on relay *B* to disconnect the rotor current above a given value of voltage. In spite of the rather large error margin in current (voltage) at which conventional relays operate, this safety circuit has the great advantage of being very simple and fully independent of the power supply and other circuits. The delay network in the circuit of relay *C* ensures that the 'start' push-button cannot be effective unless the power supply is properly warmed up.

Performance

OPERATION CHARACTERISTICS

The characteristics of the controlled motor operation with n_{sy} = 5000 rev/min is shown in Fig. 6. The discontinuity *X* indicates the instant at which the amplitudes of the pulses *n* have become large enough to trigger the Schmitt type circuit V_2 and the control system starts to operate.

SYSTEM STABILITY

Mains: possible variations from 150 to 240V after the motor has been synchronized at any specified value of f_r.

Load: synchronous operation up to additional loading torque of M_L = 6 cm.kg. (Load tests were performed by connecting resistors of known values across the tachogenerator terminals.)

Possible Applications and Developments

The control system here described is suitable for stabilizing the speed of any rotating system driven by a motor, the speed of which varies sufficiently with the voltage (current) applied to its terminals. The measure of the controllable power varies of course, with the kind of output amplifier used. For very low speeds, pulses *n* have to be amplified to a greater extent.

In view of industrial applications, it might be reasonable to redesign the control system and use semiconductors throughout which seems to be possible without any essential modification of the design principle.

Acknowledgment

Thanks are due to Professor L. Pál and to scientific division leader F. Szabó for their stimulating interest in the present work, to L. Bollók for designing the magnetic amplifier used and to G. Zsigmond, designer of the motor and selector, for useful consultation.

Distortion and Power Output of Pulse Duration Modulated Amplifiers

By E. C. Bell*, B.Sc.(Eng), A.M.I.E.E., and T. Sergent*, B.Sc.(Eng), A.M.I.E.E.

Pulse width modulation techniques are examined from the viewpoint of their application to linear high efficiency audio amplifiers. The spectrum of the modulated signal is analysed and results presented to show the distortion that may be expected and how this may be minimized by suitable choice of sampling frequency.

It is shown that with an audio bandwidth of 15kc/s, and single edge modulation, a sampling frequency of not less than 90kc/s is necessary if the output should contain no components of amplitude greater than 1 per cent of the audio signal. With double edge modulation, the sampling frequency reduces to 75kc/s.

It is further shown that in practice a power output of the order of 20W would be the maximum which would be economically available with readily obtainable transistors.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

PULSE duration modulation methods have in the main been confined to applications in the communication field. Interest has been shown, however, in the application of pulse techniques in the field of high power amplifiers[1,2,3]. The operation of amplifying devices in the linear regime is inherently accompanied by low efficiency, due to the relatively large power dissipation in the device. Any approach towards the device conditions which give lowest internal dissipation, i.e. current high and device voltage low, or vice versa, inherently increases the distortion produced.

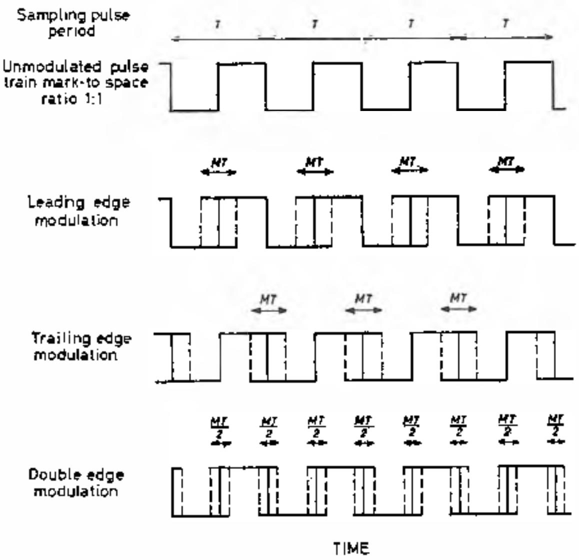


Fig. 1. Systems of pulse duration modulation

The availability of semiconductor devices, with their low collector leakage current in the 'off' condition and their low saturation voltage in the 'on' condition, makes high efficiency switching techniques even more attractive. The aim of this article is to present the modulation principles involved, to give a picture of the power and voltage spectrum of a pulse duration modulated wave and to discuss the design parameters of high quality power amplifiers using this system.

Sampling

The high power efficiency of the modulated pulse amplifier is obtained by virtue of the fact that the amplifying device handles a signal of constant amplitude, the device may therefore be either fully 'on', i.e. heavily conducting, or fully 'off', i.e. non-conducting, both these conditions are conditions of low dissipation. High energy dissipation will only occur therefore during the time when the device is switching from one condition to the other. The switching times must be designed to be as low as possible.

Information is carried by the constant amplitude signal in the form of modulation of the time duration of each constant amplitude sampling pulse. The duration is varied linearly in proportion to a sample of the amplitude of the input signal, the amplitude of the signal being sampled at intervals corresponding to the sampling period. Two basic forms of sampling have been described[4] called uniform or periodic sampling and natural sampling. Uniform sampling implies that the samples are the magnitude of the modulating signal at uniform recurrent intervals of time, i.e. the sample is taken at the same instant in the sampling period for each period. With natural sampling, the samples are taken at random times during the sampling period, the actual instant usually being dependent upon the shape of the modulation function.

Of the two types of sampling, uniform sampling will in general require the more complex modulating circuit in order to specify the sampling instant completely. Moreover, analysis shows that by this method[4], the modulated pulse spectrum contains, in addition to the required modulating signal, harmonics of the modulating signal. In this article, therefore, treatment of the modulation process will be confined to the natural sampling technique.

Systems of Modulation

Using any type of sampling technique, there are three basic ways in which the modulation may be applied to the pulse train.

The basic definition implies that the duration of the pulse must vary linearly with the modulation sample.

(a) *Leading edge modulation*

The trailing edge of the pulse occurs at recurrent

SYMBOLS

M	= modulation index, i.e. depth of modulation
ω_m	= angular modulating signal frequency
ω_s	= angular sampling frequency
p, q	= integers
$J_q(x)$	= Bessel function of the first kind, modulus x

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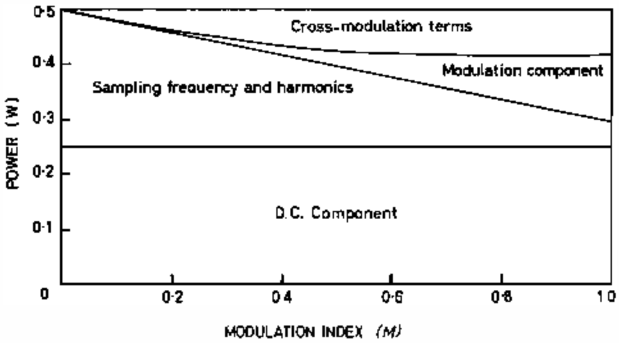


Fig. 2. Energy distribution between modulated wave components

intervals of fixed duration, the modulation being applied to the leading edge of the pulse.

(b) *Trailing edge modulation*

The leading edge of the pulse occurs at recurrent intervals of fixed duration, the modulation being applied to the trailing edge of the pulse.

(c) *Double edge modulation*

The centre of the pulse occurs at recurrent intervals of fixed duration, the modulation being applied to both leading and trailing edges of the pulse.

These modulation methods are illustrated in Fig. 1.

General Form of the Spectrum

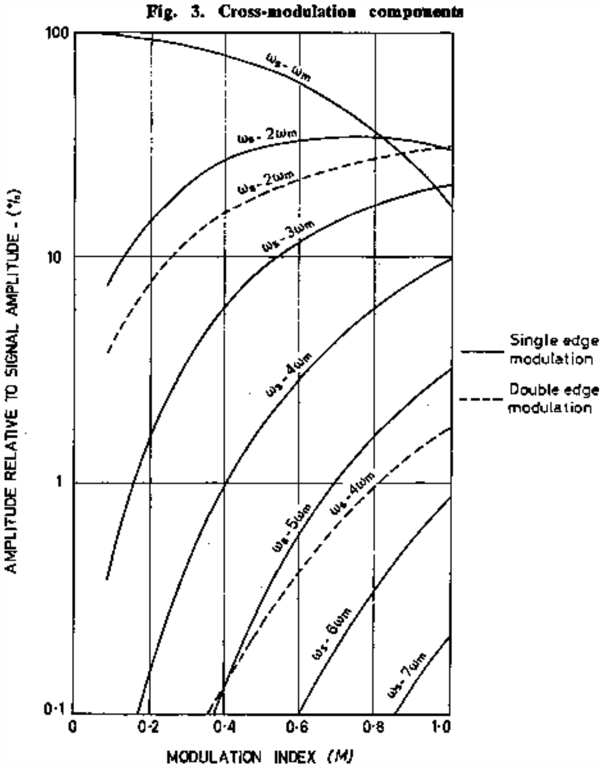
The spectra of trailing edge and double edge modulated waves have been the subject of previous published work[4] and are given below for voltage waves normalized to one volt in amplitude and of unmodulated mark-to-space ratio of unity.

(a) *Trailing or leading edge modulation*—natural sampling.

The spectrum is given below for trailing edge modulation.

$$F(t) = \tfrac{1}{2} + (M/2)\cos\omega_m t$$
$$+ \sum_{p=1}^{\infty} 1/p\pi \{\sin\omega_s t - J_0(p\pi M)\sin(\omega_s t - p\pi)\}$$

Fig. 3. Cross-modulation components



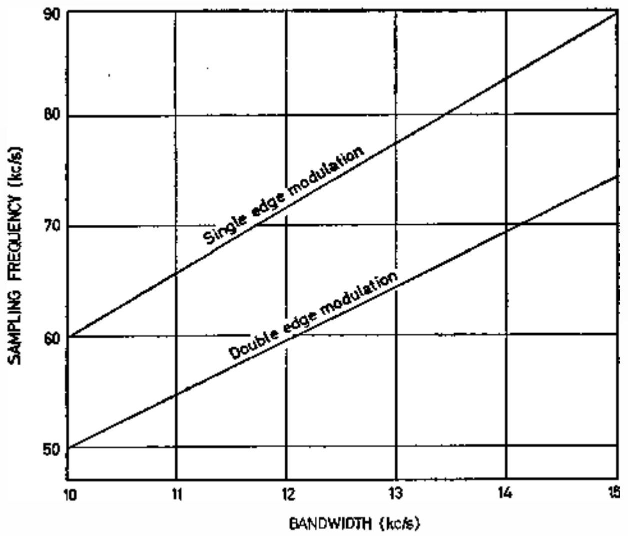


Fig. 4. Variation of sampling frequency with bandwidth on the criterion of eliminating all cross-modulation components of greater amplitude than 1 per cent

$$- \sum_{p=1}^{\infty} \sum_{q=\pm 1}^{\pm\infty} \left\{ \frac{J_q(p\pi M)}{p\pi} \sin\left(p\omega_s t + q\omega_m t - \frac{(2p+q)}{2}\pi\right)\right\}$$

The spectrum for leading edge modulation is of the same form with $-t$ substituted for $+t$.

(b) *Double edge modulation*—natural sampling.

$$F(t) = \tfrac{1}{2} + (M/2)\cos\omega_m t$$
$$+ \sum_{p=1}^{\infty} \frac{2J_0(p\pi M/2)}{p\pi} \sin(p\pi/2)\cos p\omega_s t$$
$$+ \sum_{p=1}^{\infty} \sum_{q=\pm 1}^{\pm\infty} \frac{2J_0(p\pi M/2)}{p\pi} \sin(p+q)\pi/2\cos(p\omega_s t + q\omega_m t)$$

It will be seen from both forms of modulation that each spectrum contains in the first two terms a d.c. component and a component of the modulating signal. The third term in both cases is the sampling frequency and its higher harmonics. The last term represents cross-modulation products between the modulating signal and the sampling frequency components given by such frequency terms as $p\omega_s \pm q\omega_m$ where p and q are integers.

The energy associated with each of these components has been computed for each form of modulation and is given in Fig. 2 as a function of modulation index. The energy spectrum is plotted in a normalized form in terms of a 1V peak modulated wave train applied across a 1Ω resistor.

It is interesting to note that the division of energy between the various components, i.e. d.c., sampling frequency and its harmonics, modulating frequency, and the cross-modulation products, is the same for all three forms of modulation.

Demodulation

Since a term containing the modulation signal occurs directly in the spectrum a convenient method of demodulation is to apply the pulse duration modulated waveform to the load via a low pass filter of cut-off frequency such as to prevent the dissipation of significant power in the load at frequencies above the modulation pass-band.

The Cross-Modulation Products

The angular frequencies of the cross-modulation products between the modulating signal and the sampling frequency and its harmonics, given by $\omega_s - q\omega_m$, may have substantial amplitude within the frequency range

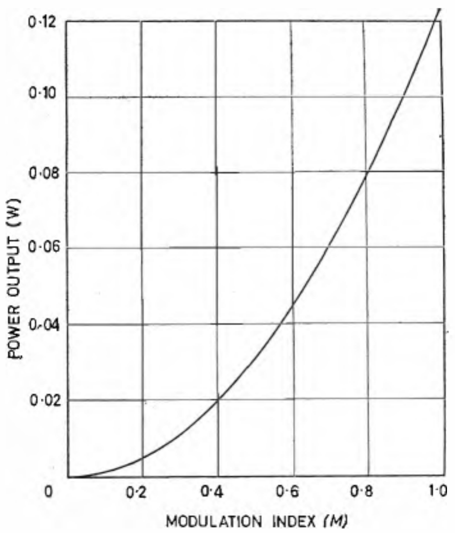


Fig. 5. Modulating frequency output normalized for 1V peak-to-peak pulse wave and 1Ω load resistor

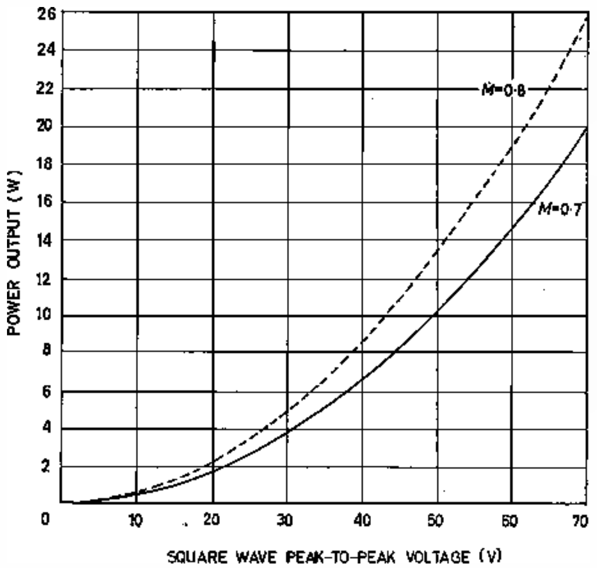


Fig. 6. Audio power output into 15Ω resistive load as a function of square wave voltage

encompassed by the modulating signal, producing distortion in the modulation signal pass-band. This would also be true of components such as $2\omega_s - q\omega_m$, $3\omega_s - q\omega_m$, etc., but as the integer p increases, the value of q required to bring the product into the modulating signal frequency range will be much higher and the amplitude will be greatly reduced.

Values of the cross-modulation products as a percentage of the signal component $(M/2)\cos\omega_m t$ when $p = 1$ are plotted in Fig. 3 for both forms of modulation as a function of modulation index.

It is interesting to note that although as described in the previous section the total energy associated with each group of components is the same for each form of modulation, the actual distribution of the energy with frequency is not the same.

This is to be expected from a consideration of the cross-modulation term in the equation for double edge modulation. It will be seen that the term involves the factor $\sin(p+q)\pi/2$. Hence when $(p+q)$ is even, the term reduces to zero. This results in alternate cross-modulation products disappearing; in particular when $p = 1$ the products given by q odd are not present as illustrated in Fig. 3.

Further it will be noted that the products remaining, given when q is even are reduced in comparison with similar terms for single edge modulation.

Optimization of Switching Frequency

The frequency interval between successive cross-modulation components is equal to the modulation frequency. At low modulation frequencies these components are grouped closely about each sampling frequency component; the distortion produced in the modulation signal pass-band is therefore inherently low. The worst case will occur when the modulating signal contains a component of maximum amplitude at the high frequency end of its pass-band. Choice of the sampling frequency will in general be determined by this case.

Referring to Fig. 3, if the system output must contain no distortion components within the modulation bandwidth of greater amplitude than 1 per cent, say, of the modulation frequency output then it will be seen for single edge modulation the first five components must lie outside the modulation pass-band. In the case of double edge modulation, only the second and fourth components are of significant amplitude.

The sampling frequency to achieve this distortion level is given by $\omega_s - q\omega_{m(max)} = \omega_{m(max)}$, hence $\omega_s = (q+1)\omega_{m(max)}$, i.e. for single edge modulation $q = 5$ and $\omega_s = 6\omega_{m(max)}$, while for double edge modulation $q = 4$ and $\omega_s = 5\omega_{m(max)}$. These results are illustrated in Fig. 4 for an amplifier of upper pass-band frequency in the range 10kc/s to 15kc/s.

The Modulating Frequency Output

For a modulated square wave of peak-to-peak amplitude V volts, the peak value of the modulated frequency output is dependent upon the modulation index M and is given by:

$$\text{Peak value} = MV/2 \text{ volts}$$

The power dissipation at this frequency in a load of resistance R ohms $= M^2V^2/8R$ watts.

This is plotted in normalized form, as a function of modulation index in Fig. 5. The graph is plotted for a square wave of peak-to-peak voltage 1V and for a load resistor of 1Ω.

For an audio amplifier design, the nominal load resistor would be typically 15Ω.

The power output into a load of this value is plotted in Fig. 6 as a function of switching voltage amplitude. It can be seen that with a depth of modulation of $M = 0{\cdot}8$, an output voltage swing of 60V peak-to-peak is required to give an audio output of approaching 20W. This voltage swing would present the maximum available with the normal transistors readily available at the present time.

Power output could be increased by using a greater depth of modulation; difficulty may be experienced in doing this with high current transistors, especially if the load is appreciably inductive.

Acknowledgments

The authors wish to acknowledge the assistance provided by the Institute of Technology, Bradford, where this work was carried out, and by other colleagues for their helpful discussion and comment.

REFERENCES

1. Ettinger, G. M., Cooper, B. J. The Design of High Power Switched Transistor Amplifiers. *Proc. Instn. Elect. Engrs. 106B*, 1285 (1959).
2. Miller, C. H. High Efficiency Amplification Using Width Modulated Pulses. *Proc. Instn. Radio Electronic Engrs. (Australia)*. (May 1964).
3. Britt, D. R. Modulated Pulse A.F. Amplifiers. *Wireless World, 69*, 76 (1963).
4. Black, H. S. Modulation Theory. (D. Van Nostrand Co. Inc. 1953).

A Reciprocal Time-Interval Display Using Transistor Circuits

By R. H. Kay*

The instrument described was developed for neurophysiological use and permits the recording of 'instantaneous' repetition frequency, in addition to recording each pulse as it occurs. The output is displayed on a c.r.t. and a linear display of pulse repetition rate is given from about 7 to 700sec^{-1}.

(Voir page 565 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 572)

WHEN observing electrical activity in living excitable tissues, it is sometimes convenient to be able to record the 'instantaneous' repetition frequency of the pulsatile action potentials directly, in addition to recording each pulse as it occurs.

Most commercial frequency meters have a minimal integration time which is too long to be useful when the impulse rate is changing rapidly. The speediest possible display is one which, when triggered by a succession of pulsed signals, measures the time interval between each pulse and its immediate predecessor and at once displays a potential difference proportional to the reciprocal of that time interval. In this way, the p.r.f. can be recorded so that not only are slow changes measured but also any rapid variations in the mean p.r.f. are not hidden by being smoothed over some finite integration time.

Probably the best circuits recently developed for neurophysiological work are those of Matthews[1] and of Huxley & Pascoe[2]; the former, though a little more elaborate, being possibly the better suited to calculable operation over a wide frequency range and the more easily tested for individual component failure.

The principles of operation of the instrument described here are exactly similar to those published by Matthews[1] and others, e.g. MacKay[3], but the vacuum tube circuits used hitherto have been replaced by functionally similar transistor circuits and the reference potentials needed in the discharge circuits have been fixed by Zener diodes: the discharge 'timing' resistors then become independent of the impedances of the potentiometer circuits which were used in previous designs to determine those reference potentials.

Circuits and Their Operation

Each input signal initiates the rapid charging of the timing capacitor (Fig. 1) to a constant potential (after a short delay equal to the duration of the brightening pulse, below). The capacitance is held at this fixed potential for a time which, added to the delay time, determines the minimum pulse interval (maximum p.r.f.) to be recorded. The potential is then allowed to decay hyperbolically with time. The capacitor is discharged through a network of resistors and silicon diodes (MR_4 *to* MR_{13}). The values of each resistance and of the potentials at which each ceases to contribute to discharging the capacitor (determined by Zener diodes MRZ_5 to MRZ_{14}) are chosen so that at any time the total discharge current is proportional to the square of the capacitor's instantaneous potential (c.f. Andrew & Roberts[4]). This potential has then a magnitude at any time inversely proportional to the time elapsed since the initial charging signal and it is displayed as a Y-deflexion on a cathode-ray tube. Arranging that the next input signal brightens the, normally dim, cathode-ray tube spot (and also initiates the recharging of the capacitor after the brightening pulse ends) effects a display in which bright spots continually signal the reciprocal of the time interval between successive pulses by the vertical displacement of the spots from the base line for zero input (Fig. 2).

Adjustment and Performance

The circuits are generally conventional. The use of Zener diodes in the timing circuits of the delay (VT_7VT_8) and pulse generator (VT_{10}, VT_{17}, VT_{18}, VT_{19}) flip-flop circuits is a modification developed in this laboratory and fully described by Bannister[5] who also discusses the complementary Schmitt trigger circuit used here (VT_3, VT_4).

Transistor VT_{24} is most of the time fully on but is cut off during the brightening pulse. In this way, the brightened spot which would otherwise continue to fall rapidly towards the base line along its hyperbolic decay curve at high p.r.f., even during the short duration of the brightening pulse, is held at a steady Y deflexion, while the brightening pulse is applied.

In Fig. 1 the input circuit is arranged to provide trigger signals at suitable d.c. levels to operate the instrument from positive going action potentials of amplitude 5 to 100V taken direct from the Y deflector plates of a Tectronix 502A oscilloscope to input 1. These signals correspond with Y signal deflexions from a few millimetres to a few centimetres. Input 2 is used for calibration and checking, e.g. by sine or square waves from an audio oscillator or by positive going pulses from a crystal-controlled time mark generator (Kay[6]). The intermediate grounded position of the input switch allows a quick display of the position of the base line for zero input to be selected when either calibrating or recording.

The preset potentiometer RV_8 holds VT_4 base at about 0·3V negative to its emitter, to keep VT_4 (and VT_3) of

TABLE 1
Transistors and Diodes for Fig. 1

Transistors:	VT_1; VT_2; VT_4—VT_{13}; VT_{18}—VT_{20}. Fairchild C 111.
	VT_{14}—VT_{16}; VT_{21}; VT_{22}. Ferranti ZT66.
	VT_{17}; VT_{24}. Ferranti ZT90.
	VT_3; OC700A; VT_{23} OC703A. Brush Clevite.
Diodes:	MR_1; MR_2; MR_4—MR_{14}. Hughes International HS 1101
	MR_3; Radiospares. Rec 50A.
Zener Diodes:	MRZ_1, MRZ_4, 4·7V; MRZ_2, MRZ_3, MRZ_5, 6·2V; MRZ_6, 8·2V; MRZ_7, MRZ_9, MRZ_{10}, MRZ_{13}, MRZ_{14}, 10V; MRZ_8, 5·6V; MRZ_{11}, MRZ_{12}, 3·3V; MRZ_{14}, 5·1V.
	Ferranti KSA Series.

* *The University Laboratory of Physiology, Oxford.*

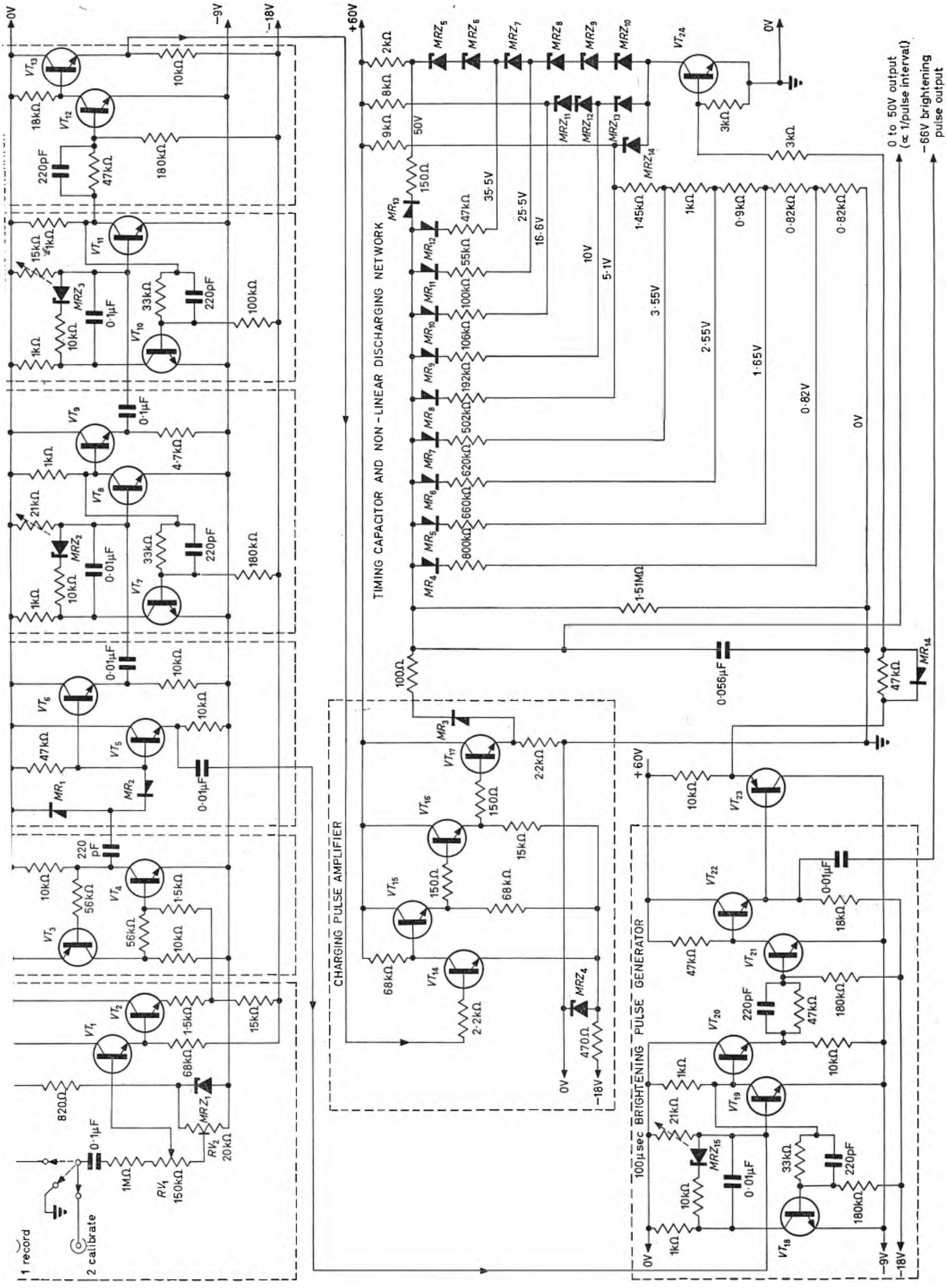


Fig. 1. Reciprocal pulse interval meter *(see Table I)*

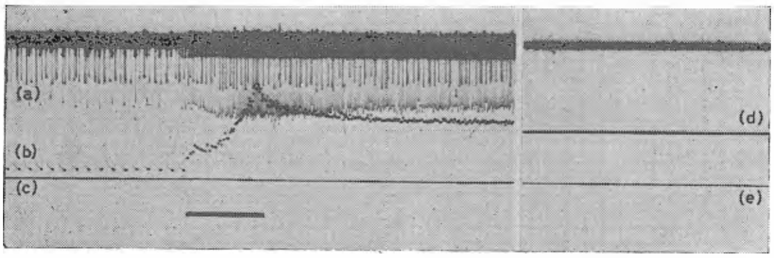


Fig. 2. The reciprocal time-interval display in operation

Left: Recording on moving photographic paper of (a) action potentials from a nerve filament containing three active fibres; the important signal, the largest of the three, being from a stretch-sensitive receptor in muscle, (b) output of the reciprocal time-interval display (c) base line for zero frequency. The muscle is extended from its original length by a linear stretch applied for the time marked by the bar then is maintained at the new length. The reciprocal pulse interval at the original length is about 14sec^{-1} and at the extended length 109sec^{-1}. The considerable overshoot to about 170sec^{-1} at the end of the stretch is a measure of a dynamic sensitivity in the receptor superimposed upon its static signalling of maintained length.

Right: A calibration (d) at 100sec^{-1}, (e) at zero frequency.

the complementary Schmitt trigger cut off until an approx. 0·5V positive-going signal arrives at VT_4 base. Potentiometer RV_1 is adjusted to suit the incoming signal amplitude and can be set to discriminate against random noise or other unwanted signals smaller than the important one (Fig. 2).

The timing capacitance and its associated non-linear discharge network are chosen for a linear display of pulse repetition rate from about 7 to 700sec^{-1}. The capacitance charging potential is fixed at 50V and this is the voltage output for maximum reciprocal pulse interval (700sec^{-1}). The d.c. amplifier monitoring this output must have an impedance large compared with 1·5MΩ, the final discharge resistor (Fig. 1). The ÷10, 10MΩ impedance probe supplied with the Tektronix 502A is used so that the total potential change available at the oscilloscope input is 5V over the frequency range. Deflexion sensitivity is varied by the oscilloscope Y gain control; zero can be offset by the oscilloscope Y shift.

The instrument has now been in operation for six months and has needed no adjustment in that time. The voltage sensitivity can be improved for other uses by feeding RV_1 from a low impedance source directly.

Acknowledgment

The author is grateful to Dr. Wu Chien-Ping who supplied the record in Fig. 2 from his own experimental work.

REFERENCES

1. MATTHEWS, P. B. C. Apparatus for Studying the Response of Muscle Spindles to Stretching. *J. Physiol. 169*, 58P (1963).
2. HUXLEY, A. F., PASCOE, J. E. Reciprocal Time Interval Display Unit. *J. Physiol. 167*, 40P (1963).
3. MACKAY, R. S. Instantaneous Frequency Measurement. *Med. Electron. & Biol. Eng. 1*, 79 (1963).
4. ANDREW, A. M., ROBERTS, T. O M. A Pulse Interval Meter. *Electronic Engng. 26*, 469, 543 (1954).
5. BANNISTER, W. J. A Transistor Delay and Pulse Generator. *Electronic Engng. 35*, 746 (1963).
6. KAY, R. H. A Time Marker for Electrophysiology. *Electronic Engng. 28*, 452 (1956).

An Interpolated Data and Speech Transmission System

Standard Telecommunication Laboratories Ltd has developed a method of using the silent gaps in a telephone conversation to send coded information without the callers being made aware. The information might be a telegraph message or digital information for data processing.

The system has been designed for use with 4-wire privately leased telephone lines for which, according to STL measurements, 68 per cent of the total 'busy time' in any one direction is not utilized. This is mainly due to the natural interchange of talking and listening roles, i.e. 50 per cent of the busy time in one direction is silence—but extensive tests have also shown that a further 18 per cent is attributable to gaps and pauses in conversation. By using IDAST (Interpolated Data and Speech Transmission system) all but a few per cent of this unused time can be filled with data, and the circuit transmission efficiency thereby improved to about 96 per cent.

In order to be able to observe and act upon forthcoming speech/silence conditions, the input speech is delayed up to a maximum of 1/5sec using a magnetic drum, while at the same time pulses are produced from timing circuits marking the start and finish of speech periods, but delayed by a similar amount to the speech itself. At the same time use can be made of the delayed pulses to switch data to the line only if the silent period exceeds the chosen minimum.

Analysis of speech shows that 90 per cent of the available silent time is in gaps having a duration greater than 0·5sec; although there is a large number of small gaps, they contribute little to the total silent time. Consequently, switching is not as rapid as a first examination of speech seems to indicate.

The minimum gap length is set by the minimum usable data burst which depends on the coding in use; the upper limit for this gap is determined by the allowable delay of speech over the system. With a CCITT maximum of 250msec for one way circuit delay and allowing 50msec for the circuit itself, the drum delay may therefore not exceed 200msec. For most data block sizes, and allowing safety margins, the optimum lies between 80 and 150msec, so that user annoyance due to delay is not a serious problem.

Input speech in the transmit direction passes first through a band-pass filter to remove unwanted noise outside the normal telephone bandwidth and a constant volume amplifier to remove average-level subscriber variations. The speech delay is obtained by a magnetic recording drum having record, replay and erase heads placed at suitable intervals around the periphery. A switch unit following the replay amplifier is operated at appropriate times by the data/speech switching logic to allow speech to pass through to the line amplifier and thus out to the transmit line.

The data/speech logic and switching circuits are controlled by the speech detector and timing pulses are produced which just 'bracket' the delayed speech and are used to operate the data/speech switching control and close the switch unit in the speech path for the required time.

A complementary output of the data/speech switching controls the connexion of the data circuit to the line output via the appropriate switch unit. Further outputs pass command signals to the data source via the data source control circuits. (Note: since the data transmission is intermittent, a buffer store following the data source is necessary.)

A further switch unit controls a signalling oscillator so that a tone is sent down the line whenever speech is transmitted to provide the appropriate synchronization at the receiving end, where signals from the line amplifier pass in parallel to switch units controlling the data and speech channels. The switch units are controlled by a signalling detector which is tuned to the tone from the transmit terminals. A band-stop filter subsequently removes the tone from the speech.

The absence of signal tone will indicate the 'data' condition and the data switch unit will operate, passing signals out on the data line. An output from the signal detector also passes to the data control circuits which can be arranged to generate any command signals required by the data receiving equipment.

LETTERS TO THE EDITOR

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

A Direct Coupled N.I.C.

DEAR SIR,—I would like to refer to your April edition, in particular to a letter by D. P. Franklin describing a directly coupled negative impedance convertor. In my opinion the system is unstable for any practical application because it contains a positive feedback amplifier whereas a current inversion negative impedance convertor is essentially a negative feedback circuit. In fact it is nothing else but a voltage controlled voltage source having a gain equal to $(Z_L + R_2)/Z_L$ connected in series to a resistance R_1 as shown in Fig. 1. In other words an amplifier with parallel derived and series fed negative feedback, having a feedback factor

$$\beta = Z_L/(Z_L + R_2).$$

The input impedance of this circuit is $-Z_L$ when $R_1 = R_2$.

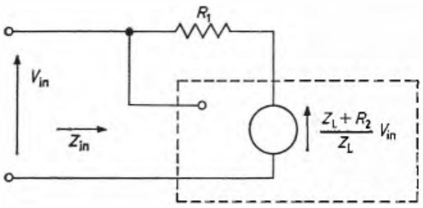


Fig. 1. Ideal n.i.c. circuit (dotted lines enclose a voltage controlled voltage source)

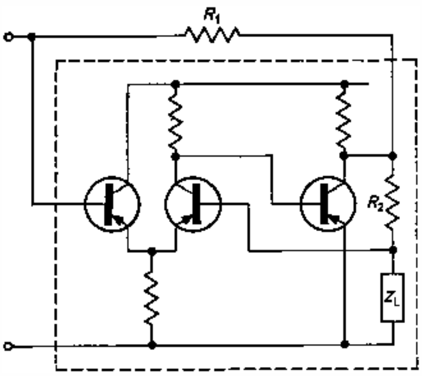


Fig. 2. Franklin's circuit (dotted lines enclose positive feedback amplifier)

Fig. 3. Practical n.i.c. circuit (dotted lines enclose negative feedback amplifier)

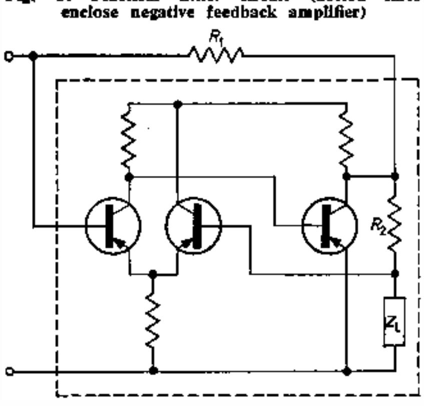


Fig. 2 shows the bare essentials of Franklin's circuit. Dotted lines enclose a positive feedback amplifier.

The idea of using a long-tail pair is however attractive from two points of view:

(a) A lower h.t. supply may be used.

(b) If the input impedance of the one transistor appears across the input terminals and the input impedance of the other transistor across the output terminals of the negative impedance convertor, then the two impedances being of the same magnitude they compensate for each other and lower values of g_{11} (= input admittance which should be zero) result.

A possible practical circuit is shown in Fig. 3. The circuit now uses a negative feedback amplifier and the discrepancy of Franklin's circuit is eliminated.

Yours faithfully,

A. ANTONIOU,

Battersea College of Technology.

The correspondent replies:

DEAR SIR,—The question of whether a circuit is stable or not cannot adequately be discussed until all the circuit impedances are defined. An isolated $-R$, if not 'held down' by series positive resistance $> R$, would certainly not settle to a stable voltage.

In a practical n.i.c. the stability will similarly depend upon the admittances connected to the respective ports, as touched upon originally at the end of my second paragraph. Going from his Fig. 2 to Fig. 3, Mr. Antoniou interchanges the circuit ports, a process which, performed upon an ideal n.i.c., will leave its behaviour unchanged. Fig. 1 is imperfect as an equivalent circuit because it does not show the terminals to which Z_L normally connects, so no conclusions can be drawn concerning this circuit with the transposed connexion.

Turning to the original diagram, it will be seen that the feedback applied at C is such as to tend to maintain the voltages at A and B equal (e.g. if $R_2 = R_6$, equal currents would be required, flowing both towards or both away from C). Whether this feedback is positive or negative depends upon the port into which the test perturbation current is injected (or across which a test voltage is applied) and, of course, the termination of the other port.

For the simple case in which R_A and R_B are connected to A and B respectively, for instance, the circuit will remain stable if

$$R_B - R_A < R_6/(G-1)$$

It follows that both R_A and R_B can be short-circuits or, with suitable R_2 and R_6, open-circuits, and that B may be short-circuited while A is open, though not vice versa.

Yours faithfully,

D. P. FRANKLIN,

EMI Electronics Ltd.

A Step-by-Step Design of a Two Way Binary Counter

DEAR SIR,—In the article on Logic Design by D. Zissos, published in the May 1965 issue, the claim is made that the final designs are independent of the switching speeds of the individual logic modules. This is a false claim—for instance the binary counter even with module 5 output connected to module 3 input still contains a hazard resulting from the speed differentials of modules 3 and 4.

This hazard is critical when the circuit is in the transition:

Module	1	2	3	4	5	6
$X = 0$	1	0	0	1	0	0
to $X = 1$	1	0	0	0	0	1

If module 4 responds to X going to a 1 before module 3 does so, then module 4 going to a 0 can put a temporary false 1 in module 3 which changes the state of the bistable formed by modules 1 and 2. The response of module 3 to X going to a 1 cannot now remove this change of modules 1 and 2, and the circuit now becomes stable in the state:

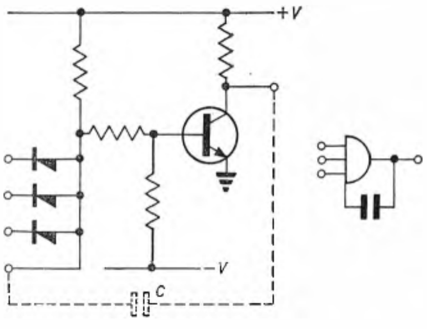
Module	1	2	3	4	5	6
$X = 1$	0	1	0	0	1	0

This is an incorrect cycle and counting does not occur.

A manner in which speed differential between modules 3 and 4 may occur is if the drive x is a sine wave of very slow periodicity with respect to the logic speed of propagation of the modules and if the module circuit design is such that they may have a large and possibly varying input threshold. In these circumstances X may effect modules 4, 3, 1 and 2 in that order before it effects module 3 and the circuit does not count.

I came across this feature when I attempted to use this logic configuration to count directly off a 50c/s sine wave in order to avoid the necessity of squaring with a Schmitt trigger circuit before using a more conventional 'capacitor steering gate' type of counter. The circuit did not count at all, and I finally solved the problem by putting a Miller capacitor across modules 3 and 4 calculated to delay their outputs

Fig. 1. Use of Miller capacitor



longer than the time the sine wave took to scan the maximum voltage threshold of the module input.

This approach while destroying the true d.c. sequential design of the circuit gave an entirely calculable design and the bench model worked with the capacitors and stopped when they were taken off.

Yours faithfully,

T. H. BALDWIN,

Widnes, Lancashire.

The author replies:

DEAR SIR,—The claim that the final circuit is independent of time delays is made only on the assumption that the output of module 5 is connected to module 3. This is obvious from the time diagram shown in Fig. 1, where one module only switches at a time.

Such a time diagram can only be drawn on the assumption that a primary signal will always affect a module it is connected to before the other circuit signals feeding the same module. This assumption is justified in the case of a square wave input.

If, however, the input is not a true square wave, this assumption cannot be made, and as Mr. Baldwin suggests, it is possible for a secondary signal to affect a module before the primary signal, resulting in a static hazard which may in turn affect the correct operation of the circuit. In such a case the method of eliminating static hazards explained in the article can be used to make the circuit independent of relative switching speeds of the modules.

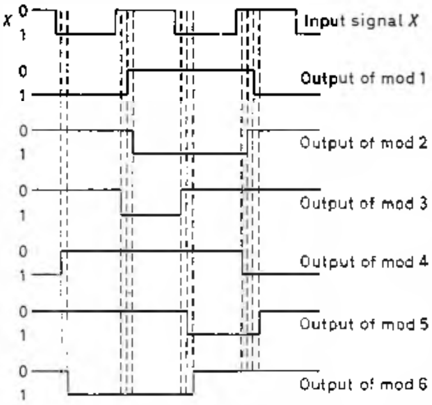
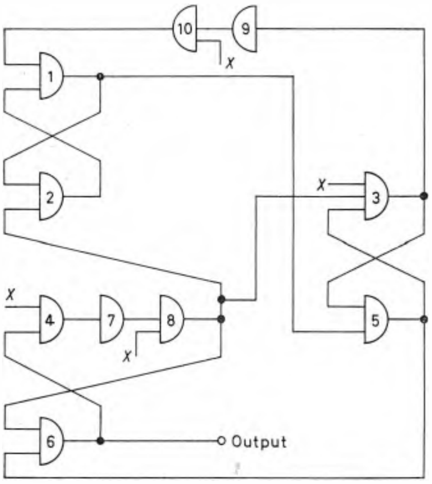


Fig. 1. Time diagram of two-way counter with output of module 5 connected to input of module 3

Fig. 2. Use of gates



The static hazard appearing at the output of module 3 which Mr. Baldwin uses to support his case can readily be prevented from reaching module 1 by introducing the appropriate gate between the output of module 3 and the input of module 1, which is closed when $X = 1$.

Similarly a static hazard will appear at the output of module 4, if X becoming 1 affects module 3 before module 4. A similar gate at the output of 4, which is closed when $X = 1$, renders the circuit independent of relative propagation times.

The circuit shown in Fig. 2 worked satisfactorily for sine wave inputs as low as 10c/s. It must however be pointed out that these additional gates are only necessary when the input is a sine wave of low frequency.

A different solution to this problem is to redesign the circuit, so that the primary signal feeds one module only. Such a circuit of course will not necessarily be minimal. Research is at present being carried out at the College on this problem.

D. ZISSOS,

City of Liverpool College of Technology.

Feedback Transistor Oscillators

DEAR SIR,—I would like to suggest that in the correspondence in your June issue arising out of the article by B. Shahzadi on the analysis of the transistor Colpitts oscillator, some confusion has arisen because of the use of the expressions common emitter, etc., with reference to oscillators. A common emitter circuit is one in which the emitter is common to input and output, but an oscillator has no input, so that when the expressions common emitter, etc., are used with reference to oscillators they must have some other meaning. There appear to be two such meanings. These are:

(1) The terminal called common has been used as the datum node for the purpose of nodal analysis. This appears to be what Mr. Shahzadi means in the final sentence of his reply, but the choice of datum node is, of course, arbitrary, so that this tells us nothing about the circuit.

(2) There is negligible oscillation frequency voltage between the terminal called common and the supply terminals. This appears to be Mr. Munro's meaning.

I suggest that it would be better not to use the expressions common emitter, etc., of oscillator circuits, but to speak only of earthed (or grounded) emitter, etc. circuits.

Yours faithfully,

K. S. HALL,

Northampton College of Advanced Technology, London, E.C.1.

DEAR SIR,—It appears that an old fallacy which should have been killed long ago is again rearing its head.

A transistor (or valve) has three terminals. Consequently when it is acting as an amplifier it is reasonable to speak of common-emitter connexion, common-base connexion, and common-collector connexion (and to use the corresponding expressions for a valve), according to which electrode is joined to the terminal common to input and output. An oscillator has no signal input, and such expressions are therefore meaningless.

An oscillator of the type under discussion consists of a three-terminal amplifier and a three-terminal network. Most people, I think, will feel that the simple and basic way of analysing the circuit is to take C and E as the input terminals of the network, and B and E as the output terminals, Fig. A. Looked

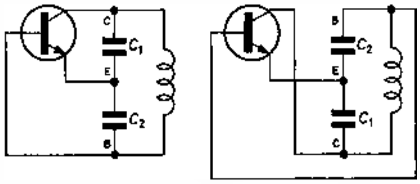


Fig. A (left), **Fig. B** (right). Arrangements of oscillator circuit

at this way the network gives a change of sign at the oscillation frequency.

But we can, of course, if we wish to, regard C and B as the output terminals, Fig. B. The network will then be said to have an output voltage in phase with the input voltage and to have a voltage gain greater than unity, and the amplifier will be said to be giving an inphase voltage gain (magnitude less than unity). The temptation to think that we now have a different oscillator, that the transistor is working in any real sense in a different way (that it is working as an emitter-follower, or anything of that sort) must, however, be resisted firmly. The oscillator is quite unchanged; all we have done is turn part of the circuit diagram round on the paper.

My own rule for avoiding falling into any such fallacies is to analyse all circuits with the emitter (or cathode) taken as the reference of potential (earthed, if you like—although earthing is a practical matter and of no theoretical significance in a self contained system). This will often show the essential identity of circuits which have been regarded as different, and disprove imagined superior properties of one or another.

It is clear, of course, that Mr. Shahzadi in his reply is quite familiar with the above; but perhaps another reply, in different language, may be helpful. An earlier treatment of the subject is given in a letter to the *Wireless Engineer*[1].

Yours respectfully,

E. F. GOOD,

Malvern, Worcestershire.

REFERENCES

1 BAXANDEL, P. J. *et al.* (letter). *Wireless Engr.* 33, 129 (1956).

BOOK REVIEWS

Electronic Amplifiers for Automatic Compensators

By D. Ye. Polonnikov. Translated from the Russian by Rita R. Inston. 323 pp. Demy 8vo. Pergamon Press. 1965. Price 80s.

THIS book deals with the application of electronic amplifiers to automatic measuring and control equipment. The successful operation of such equipment depends very largely on the adequacy and reliability of the amplifier which has to operate under conditions quite different from, and in certain respects considerably more stringent than, those for most other, more conventional uses.

The book was originally published in the U.S.S.R. in 1960 and refers largely to the intensive development of electronic circuits carried out there in the post-war period which culminated in the successful launching of the first space vehicle, Sputnik I, in 1957. Written by a member of the Russian Academy of Science, who was mainly concerned with this development, the book makes interesting reading as it conveys some of the basic ideas and the general attitude in the U.S.S.R. towards automation. Emphasis is laid throughout on the development of a relatively small number of sophisticated circuits incorporating high performance with great flexibility which can be adapted to very wide ranges of applications, including bio-engineering. The major characteristics aimed at are: extremely high input impedance, very high sensitivity, low noise, rapid response—and small overall dimensions. Unfortunately, the work described is confined to the pre-transistor era, up to 1957, and what was then called 'small' would hardly deserve that epithet now.

After an outline, in the opening chapter, of the conditions under which amplifiers for automatic compensating systems have to operate, a brief survey is given in chapter 2 of the circuits available for a.c. and d.c. operation, including those for very high input impedance. Chapter 3 then deals with the calculation of the principal parameters of amplifiers and the determination of their characteristics. The section on the experimental methods for determination of these characteristics will be found particularly useful. The next three chapters bring a detailed discussion, for the same three types, a.c., d.c., and high-sensitivity current amplifiers, of perhaps the most important aspect in the layout of compensator amplifiers, i.e., the design of the input circuit. Vibrator convertors are included in the discussion of the latter two types. The final two chapters deal in a fairly conventional manner with voltage and power amplifiers and power supplies. The bibliography cites some forty papers published in the U.S.S.R. up to 1958, and about half that number from the West.

The translation is of a high quality, modern electronic terminology current in the West being used throughout. The occasional confusion of the definite and indefinite articles can be misleading in certain circumstances; this is no doubt due to the non-existence of either 'the' or 'a' in the Russian language.

The book should be of value to electronic design engineers, particularly those concerned with automation. A new edition will be awaited with interest, especially if it extends the discussion to transistorized circuits.

E. BILLIG

The Physical Basis of Electronics

By J. G. R. van Dijck. 362 pp. Med 8vo. Centrex Publishing Co., Eindhoven. 1964. Price 55s.

ON casual inspection this appears to be a well-produced and interesting book covering the broad field of physical electronics at an elementary level. Unfortunately it soon becomes apparent that appearances, in this case, are very deceptive for apart from its attractive format and the clarity of the diagrams the book has nothing to recommend it. For example, the coverage is so wide that the treatment is quite superficial.

However, there are far more serious criticisms than this. Firstly there are more misprints than are usually found in a book of this size. Secondly there are a number of outright errors of fact such as "When a proton falls on a surface with a high energy it can disappear entirely giving place to a position plus a negaton" (p. 30). In fairness it must be said that this is probably a misprint, of which there is more than one in this sentence. The same cannot be said of the following howlers. "It will be seen that an electron with a kinetic energy of only 1eV has three times the mass of an electron at rest" (p. 44). This information is repeated in an accompanying table. We are told on p. 293 that the function of the screen grid of a tetrode is to accelerate the electrons "so that they have more chance of reaching the anode field". The author goes on to tell us that secondary electrons produced at the anode of a tetrode may pass to the screen in the opposite direction to the primary electrons "and hindering their passage". This undesirable effect may be prevented by the introduction of a suppressor grid which "is usually connected to the cathode thus slowing down the electrons accelerated by the screen grid which are by this time sure of reaching the anode". Apparently molecules in the two appropriate energy states of the ammonia maser are separated by an inhomogeneous *magnetic* field. "The gas maser has, however, the disadvantage that the excited radiation has a rather wide frequency range" (p. 194). Since the ammonia maser has a bandwidth of at best 1kc/s at 24 000Mc/s, and gas masers are used as frequency standards, this statement is surprising.

Further examples could be given but the point has been made. There are also many misleading half-truths and unqualified statements. Thus it is said that the emitter of a transistor is of n-type material, that a parallel beam of light will remain parallel no matter how long it is, that a laser is based on a single crystal of ruby and that the mirrors must be precisely the right distance apart.

Apparently the book has already been published in several other languages. It is not possible to say whether the many serious faults are peculiar to the English edition, but in its present form the book is, without doubt, quite unacceptable.

W. A. GAMBLING

Linear Analysis of Electronic Circuits

By Glenn M. Glasford. 580 pp. Med. 8vo. Addison-Wesley Publishing Co. Inc. 1965. Price 113s.

THIS book is intended to provide a fundamental course in electronic circuits at undergraduate level. As the title suggests, the text is restricted to circuit operation in the linear regime and topics such as waveform shaping, switching circuits and the like are not considered. The author has previously worked in the Radiation Laboratory at the Massachusetts Institute of Technology (M.I.T.) and with the Allan B. Du Mont Laboratories; at the present time he is Professor of Electrical Engineering at Syracuse University.

The first chapter considers the various types of electronic device both thermionic and semiconductor; little is said of the physics of device operation and the author is mainly concerned with the discussion of the equivalent circuits. The list of electronic devices is by no means complete and such topics as photoelectric cells, cathode-ray tubes, cold-cathode tubes and thyristors are not mentioned. The second chapter is simply circuit theory and much of the material in this chapter, such as the Laplace transform and Fourier series, is likely to be covered in associated courses of study. Chapter three deals with the use of valves and transistors as elements in actual circuits. The cathode-follower and emitter-follower together with the normal amplifier configurations are included. Chapter four, which represents almost one quarter of the book, covers complete amplifiers and deals with frequency response compensation, band-pass circuits and power-output stages. The final chapter is on feedback including the operational amplifier and the degenerative voltage stabilizer. A series of problems (without answers) is given in the appendix at the end of the book.

From the brief discussion of the content it will be seen that the book is mainly concerned with amplifiers and hence would only be useful for a relatively small part of an undergraduate course. This volume is fairly expensive and it does not seem to be particularly well suited for a first course of study in electronics.

V. H. ATTREE

Power Travelling Wave Tubes

By J. F. Gittins. 276 pp. Crown 4to. The English Universities Press Ltd. 1965. Price 50s.

SINCE the invention of the t.w.t. in 1943 an understanding of the interaction between moving electron streams and electromagnetic waves has developed rapidly and includes many different basic types of amplifier and oscillator.

As the title suggests, this book concerns itself principally with one facet of this subject, namely, the high power O-type amplifier. Furthermore, as the author expresses in the introduction, it is basically aimed at the engineer. Bearing in mind these restrictions the book does justice to the claims. The theory is, with reservations, covered elsewhere in the literature, but is collected here in a way which is convenient for the designer, giving him a basis for understanding the processes permitting calculation of the primary beam and circuit parameters

The book is divided into three main sections on the O-type tube with a short chapter on M-type interaction. The main sections are devoted to the slow wave structure and ancillary r.f. components, beam formation focusing and collection, and the engineering aspects of tube making. After an introductory chapter on the current state of the art there follows an interesting section on a vector model providing a physical picture of the growth phenomenon succeeded by a Pierce wave treatment for small signals with mention only of the various large signal theories. The following chapters then treat the topics outlined above.

The balance between the purely theoretical and experimental approach is quite well preserved; however, considering the author's experience on high power c.w., t.w.t.'s it is perhaps a little disappointing that more space is not devoted to the vast data gathered on, say, the clover leaf structure on which the S.E.R.L. has considerable experience. Although a large amount has already been published it is always useful, as this book purports, to have the information collected in a single reference.

A relatively minor criticism of the layout is that the size of the diagrams and the associated lettering is not always well related to the information content or to the rest of the page, also, designation of the space harmonics on the ω-β diagrams would have given better correlation with the script.

On the whole the book fulfils its aims well, is clearly written and freely illustrated on a pleasant format and containing, at first reading, very few and minor errors.

F. J. NEWLAND

Potential Barriers in Semiconductors

By B. Gossick. 153 pp. Demy 8vo. Academic Press Inc. 1964. Price 17s. 6d.

This book deals with the physical processes associated with potential barriers in semiconductors and includes such topics as energy bands, charge transport, variations in electrochemical potentials under non-equilibrium conditions and the current by electrons and holes associated with fluctuations in a potential barrier.

Automatic and Remote Control

By V. Broida, D. Barlow, O. Schafer. Vol. 1 749 pp. Vol. 2 695 pp. Demy 4to. Butterworth & Co. 1964. Price £38

The proceedings of the second congress of the International Federation of Automatic Control (I.F.A.C.) held at Basle in 1963 are contained in these two volumes.

The first volume is entitled 'Theory' and deals with non-linear, discrete, optimal and self-adjusting systems.

Some 90 papers were presented at this section of the conference.

The second volume is divided into two sections 'Applications and Components' and covers the application of automation in the electrical utility field, steel, oil and chemical industries and in industrial processes. The automatic control of aerospace systems is also dealt with.

Mechanical, hydraulic, pneumatic and electromechanical devices are contained in the components section which also deals with electronic components and their reliability, digital devices and instrumentation.

Some 80 papers were presented at the application and components section.

Variational Techniques in Electromagnetism

By L. Cairo and T. Kahan. 152 pp. Demy 8vo. Blackie & Sons Ltd. 1965. Price 27s 6d.

This monograph discusses the application of variational methods to electromagnetic wave propagation, particularly at conducting surfaces and in waveguides.

In the first three chapters the authors give a summary of some fundamental mathematics, such as the dyadic calculus and Green's function, which are required for their subsequent development of variational methods.

The translation from the French has been made by Dr. G. D. Sims, Professor of Electronics, Southampton University.

Handbook of Electron Tube and Vacuum Techniques

By F. Rosebury. 597 pp. Med. 8vo. Addison-Wesley Publishing Co. 1965. Price 132s.

The aim of this book is to supplement the basic principles and to outline procedures and techniques for the construction and manufacture of electron tubes and other evacuated devices.

It is based on the first and second editions of the Tube Laboratory Manual originally produced by the Research Laboratory of the Massachusetts Institute of Technology.

Progress in Dielectrics Volume Six

By J. Birks and J. Hart. 334 pp. Royal 8vo. Heywood Books. 1965. Price 84s.

This book is the sixth in the annual series of volumes under the editorship of Dr. J. B. Birks of Manchester University, and Prof. J. Hart of Carleton University, Ottawa, and contains five papers dealing with new aspects of dielectric phenomena.

Worked Examples in Direct Current for Engineering Students

By P. Freeman. 138 pp. Med. 8vo. Macdonald & Co. 1965. Price 30s.

This book contains about 100 worked examples in direct current as applied to resistivity, temperature co-efficient, power, energy, networks, distributors, ring mains, magnetic circuits, inductance, capacitance, generators and motors.

It is the companion volume to 'Worked Examples in Alternating Current' by the same author and is intended for students in the first year Engineering Degree course and Diploma of Technology and similar levels.

Introduction to Electronics

By Bureau of Naval Personnel. 145 pp. Med. 8vo. Constable & Co. Ltd. 1965. Price 8s.

This book was prepared by the U.S. Navy Training Publications Centre, Washington, for the Bureau of Naval Personnel and is intended for use by trainees whose duties require an elementary general knowledge of radio, radar and sonar.

Proceedings of the International Council of the Aeronautical Sciences

By T. Karman. 1175 pp. Med. 8vo. Macmillan. 1964. Price £18

The proceedings of the Third Congress of the International Council of the Aeronautical Sciences (I.C.A.S.) held at Stockholm in 1962 are published in this book. Fifty-two papers were presented at this Congress, previous congresses being held at Madrid (1958) and Zurich (1960).

Electronic Equipment in Industry

By W. Gilmour. 265 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 50s.

The treatment of the techniques and principles of electronic instrumentation and control equipment is practical and non-mathematical in this book.

The main subjects dealt with are electronic measuring techniques—both analogue and digital types: communication and telemetry; servo-control; computers; data logging; process control; phototechniques; assessing requirements for a given purpose; designing to meet those requirements; progress form 'breadboard' to production stage; and operational installation, testing and fault finding.

Elementary Circuit Properties of Transistors

By C. Searle, A. Boothroyd, E. Angelo, P. Gray, D. Pederson. 306 pp. Demy 8vo. John Wiley & Sons. 1965. Price 20s.

This is the third book in the series dealing with transistors and other semiconducting devices prepared by the Semiconductor Electronics Education Committee at the Electrical Engineering Department of the Massachusetts Institute of Technology.

Digital Instruments

By K. J. Dean. 181 pp. Crown 8vo. Chapman and Hall Ltd. 1965. Price 25s.

The fundamental principles of digital instruments are described in this book starting with an outline of information theory and binary decimal codes. It also explains how information is made available to an instrument and how its accuracy and reliability can be effected by the way in which logical processes are organized.

Correction

On page 337 of the May issue it was stated that volume 1 of the Dictionary of Electronics published by Rhode and Schwarz contains 6 000 words. This should read 60 000 words.

ELECTRONIC EQUIPMENT

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

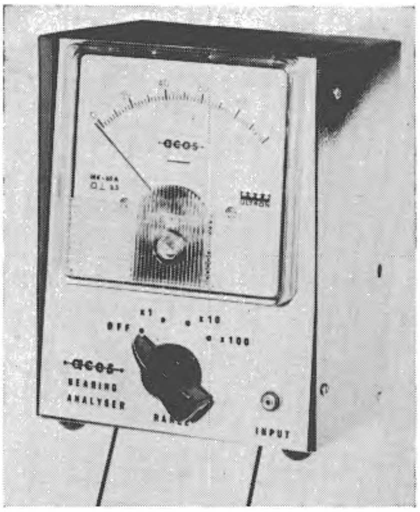
(Voir page 559 pour la traduction en français; Deutsche Übersetzung Seite 566)

BEARING ANALYSER

Cosmocord Ltd, Eleanor Cross Road, Waltham Cross, Hertfordshire

(Illustrated below)

This range of instruments has been developed to enable plant maintenance engineers to quickly determine the comparative state of the bearings etc., in machine tools. The equipment will also indicate an overload condition if the cutting tool is advanced too rapidly into the material being worked. The analysers eliminate the need to change bearings etc., at predetermined intervals and also safeguard against the production of faulty machined parts.



The equipment comprises of a sensing device usually mounted on the live bearing housing of the machine and the read-out device.

For automatic production control purposes an indicating device fitted with an overload cut-out is available, this cut-out being used to stop the process if a fault should occur, or give a suitable warning. The whole system is robustly constructed and under normal conditions virtually indestructible, as the sensing device is a small solid unit and the read-out circuit completely transistorized.

EE 84 751 *for further details*

DIGITAL VOLTMETER

Hewlett Packard, Dallas Road, Bedford

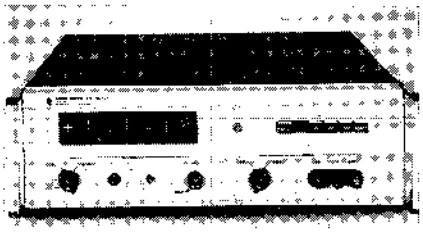
(Illustrated above right)

This new, all solid state digital voltmeter, model 3460A, offers a combination of high accuracy, noise immunity, high and constant input impedance, and high speed, through use of a new operating principle.

Read-out of the voltmeter is a full five digits, with a sixth digit for overranging. The maximum resolution on the lowest range is 10µV and the input impedance is a constant 10MΩ on all ranges. Full accuracy (±0·005 per cent of reading, ±2 counts, in environments from 10 to 40°C) is maintained even at the instrument's highest operating speed, up to 15 readings per second.

Model 3460A is both integrating and potentiometric. It is, in a sense, an automated, integrating, guarded differential voltmeter. It takes the reading in two rapid, successive sampling periods. In the first, a floated and guarded voltage-to-frequency convertor integrates the unknown, generating a series of pulses whose rate is proportional to the instantaneous input voltage. These are counted; this count actuates the most significant digits of the reading. Now the total count is stored in the counting units and the stored count is transferred to a digital-to-analogue convertor. This convertor generates an analogue signal exactly equal to the stored count (reading) and this signal is compared with the unknown. The difference is fed to the voltage-to-frequency convertor and integrated into pulses which are counted in a second sampling period. These counts are entered into the fifth and sixth digits. The whole count is now transferred to the read-out display.

The instrument is potentiometric in every sense, generating a proportional signal and comparing it with the input, relying for its accuracy primarily on precision resistance ratios and a stable reference voltage. Unlike other potentiometric meters, however, it is relatively immune from normal-mode noise, without resorting to speed-reducing input filters. Common-mode rejection, with 1kΩ unbalance, is 140dB at all frequencies.



Any of four ranges, or autoranging, may be selected by push-button. Because model 3460A has 20 per cent overranging on all ranges, it will measure up to 1 200V d.c. the highest voltage being presented with full accuracy on the six digit display. Reversible counters are used for integration of signals varying around zero.

Operation in systems has been facilitated by including binary coded digit printer output (1 - 2 - 2 - 4), remote programmability of range and measurement function and accessories for making a.c. and resistance measurements. Front panel switches transfer operation to rear terminals.

EE 84 752 *for further details*

REED LOGIC ELEMENTS

Flight Refuelling Ltd, Wimborne, Dorset

(Illustrated below)

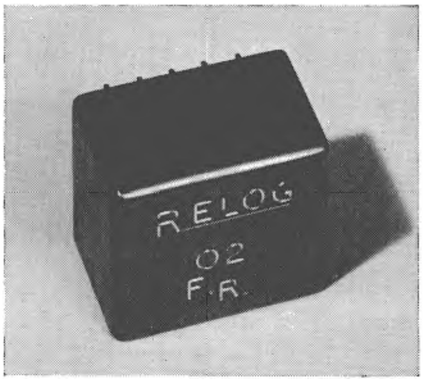
To meet the demand for simple logic switching systems combining multiple input and output, with reasonably fast switching speed at moderate cost, Flight Refuelling Ltd has developed a range of reed logic elements—'Relog'—built around the Hamlin MRC-1 miniature reed switch.

For certain types of applications, the reed logic system can be used with advantage, particularly in the machine tool and mechanical handling industries; for although its operating speed is slower than transistorized logic, in many applications in these fields it is the input and output devices that are in fact the speed limiting factors for any one complete system.

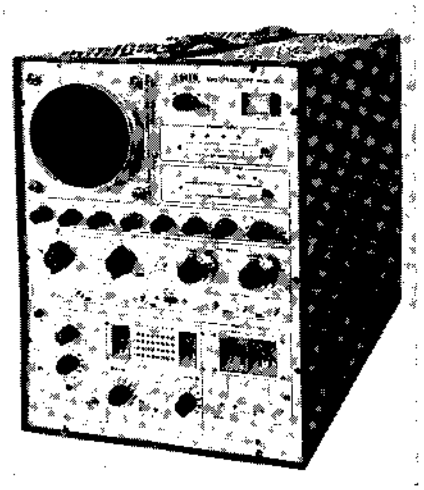
Hence the reed logic provides an adequately fast system at a lower cost in that the necessary ancillary equipment is less complicated and only one single power line is required. A package density approaching that of transistorized logic can be achieved insomuch that up to 36 'Relog' elements can be mounted on a 10in × 10in board.

Two logic functions, OR and NOR, are currently available, each as a two or four input element. Both functions have the added facility of increasing the number of inputs indefinitely by the addition of further diodes.

Due to the use of a reed switch capable of handling currents up to 0·5A, a large number of logic loads (25 maximum) can be driven from one element and complete electrical isolation of the output circuit of each element is achieved.



EE 84 753 *for further details*



100-CHANNEL ANALYSER

Distributed by Nuclear Enterprises (G.B.) Ltd, Sighthill, Edinburgh, Scotland

(Illustrated above)

The Spectrascope Mod.100-channel Analyser, the latest addition to the range of Laben nuclear spectrometry analysis systems, can now be supplied by Nuclear Enterprises (G.B.) Ltd.

This portable, accurate and flexible instrument consisting of an amplifier, convertor, memory logic, decoding for print-out and multi-scaler, operates both in addition and subtraction, and the address and content registers are accessible in series, or, on request, in parallel. It can operate in the following modes: derandomizing, multi-scaling, sampling, averaging and pulse height analysis. The analyser can be used as an independent single channel unit in multi-scaler mode, and has a built in *RC* clipped non-overloading amplifier. It can be operated with a serial printer and, on request, with a parallel printer. Specifications include: a count capacity of 10^5-1, digitizing rate of 10Mc/s, a 1·5μsec switching time, a 6μsec store cycle and a maximum total dead time of 17μsec.

A plug-in unit Mod.100A, which can be added to the basic analyser or used independently, is also now available. The Mod.100A includes a high voltage supply for nuclear detectors, a timer programmer performing preselection of measure and waiting times, and a channel selection unit for data integration and the selection of output signals to allow automatic stabilization with the external unit.

EE 84 754 *for further details*

DIGITAL VOLTMETER

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated right)

A low-cost digital voltmeter and a.c.-d.c. convertor are now available from Dawe Instruments Ltd. Designated the type 652 and type 653 respectively, they can be used separately or together to provide a four-digit read-out of a.c. or d.c. voltage to 0·1 or 0·2 per cent accuracy.

The type 652 digital voltmeter is offered in two basic versions, type 652A and type 652AX. The first uses a simple standard cell reference voltage to measure down to 0·2mV with an overall accuracy of ±0·2 per cent of maximum read-out on each range. The type 652AX uses a Zener diode reference to measure down to 0·1mV with an overall accuracy of ±0·1 per cent of maximum read-out. Both versions have four ranges giving maximum read-out of 1, 10, 100 or 1 000V d.c. on mechanical counters with automatic decimal point indication. Input impedance is 2·2MΩ, sufficient to avoid loading most circuits.

Design is based on a self-balancing servo system, the d.c. input voltage being fed to a comparator amplifier together with a d.c. reference voltage controlled by a motor-driven potentiometer. The comparator amplifier produces an a.c. output, whose amplitude and phase are related to the difference between the signal and reference levels, this output being fed to the control winding of a two-phase electric motor. The motor drives the display counters and moves the potentiometer slider until the reference voltage balances the d.c. input to the comparator.

Read-out time is an average of about 2sec, with a maximum of 5sec (for maximum read-out). Set zero facilities are incorporated. The instrument may be powered by a 50 or 60c/s mains supply and is contained in a compact portable cabinet rugged enough for use in the workshop. Dimensions are 7½ by 6 by 11in high, and weight is about 9 lb.

The type 653 a.c.-d.c. convertor facilitates the accurate measurement of a.c. voltages by means of the type 652 digital voltmeter. Its circuit consists of an input attenuator, high-gain amplifier and averaging filter output stage. Input signals may be applied to the convertor from millivolt levels to 500V r.m.s., at frequencies from 50c/s to 25kc/s. The input may be floating, or earthed, but output is always floating, providing a maximum of 1V d.c. on the lower three ranges and 0·5V d.c. on the uppermost range. Nominal accuracy is ±0·3 per cent of maximum output on each range from 50c/s to 10kc/s, and within ±0·4 per cent up to 25kc/s; linearity on all ranges is 0·05 per cent. The unit is

mains-powered and is housed in a cabinet designed to match that of the type 652.

EE 84 755 *for further details*

TELEPHONE CABLE TEST SET

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport Monmouthshire

(Illustrated above)

A portable, self-contained telephone cable test set offering distinct advantages to cable installation engineers has been announced by Standard Telephones and Cables Ltd.

Known as the 74226, the unit can rapidly measure mutual capacitance capacitance unbalance and percentage resistance unbalance. Measurements may therefore be made of the various combinations of capacitance unbalance encountered in separate cable sections having quadded or paired construction so that optimum section-joining configurations may be employed with the object of minimizing crosstalk and interference.

The measuring ranges are as follows Capacitance unbalance:

Range 1 280-0-280pF;

Range 2 1 100-0-1 100pF;

both with an accuracy of ±(1 per cent + 2pF). Mutual capacitance of up to 0·1121μF can be dealt with at an accuracy of ±50pF, and for resistance unbalance ±0·5 per cent of conductor loop resistance may be measured to an accuracy of 0·01 per cent.

Operating features include an arrangement to tilt the box for easier visibility and optional scale illumination for use in poor light. In addition the a.f. output in the headphones is limited when the bridge is out of balance so as to obviate operator discomfort.

The bridge circuit works in conjunction with a transistorized oscillator and amplifier and two versions are available having operating frequencies of 800 and 1 000c/s respectively. Power is supplied internally from two 4·5V dry batteries.

The set is housed in a wooden box having carrying handles and a detachable lid. The overall dimensions are 17 × 12 × 10in and the weight 43 lb.

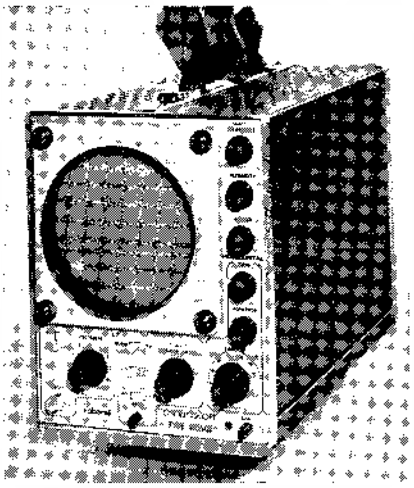
EE 84 756 *for further details*

OSCILLOSCOPE

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustrated below)

This new oscilloscope, the RO 501, has a 5in c.r.t. operating at 3·3kV



stabilized to ensure high brightness displays at any sweep speeds and uses a single printed circuit board with all main components coded.

It has a bandwidth of d.c. to 6Mc/s, a rise time of 55nsec, sensitivity 50mV/cm to 20V/cm covered in nine calibrated ranges (1, 2, 5 steps) and a measuring accuracy of ±5 per cent.

The sweep range is 1μsec/cm to 100msec/cm in six calibrated steps and a 10 : 1 continuously variable control extends the range to 1sec/cm; a ×5 variable magnifier offers a maximum sweep speed of 200nsec/cm. A versatile trigger circuit allows the trigger point to be selected and a built-in television frame integrator allows the time-base to be triggered from television frame pulses. Sensitivity: internal 5mm; external 4V peak-to-peak.

An internal calibrator is fitted, the two colour graticule is edge lit, the oscilloscope accepts standard cameras on 5in centres, and the overall size is 9¾in high, 14¾in deep and 7½in wide. It weighs 17½ lb.

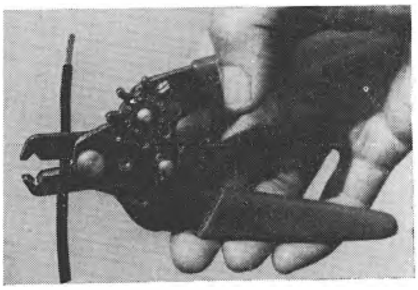
EE 84 757 *for further details*

WIRE CUTTER AND STRIPPER

Multicore Solders Ltd, Multicore Works, Maylands Avenue, Hemel Hempstead, Hertfordshire

(Illustrated below)

The new 'Bib' model eight wire



cutter and stripper is fitted with a preset selector gauge for selecting any Standard Wire Gauge between 12 and 26. By correct use of the selector, the insulation can be stripped without danger of damage to the wire. This tool is fitted with cushioned plastic handles which provide an easy grip.

EE 84 758 *for further details*

DIGITAL VOLTMETERS

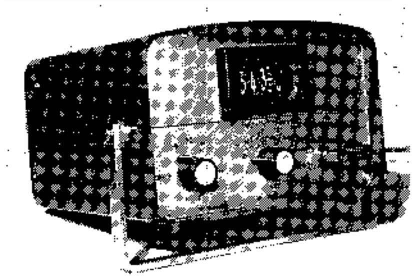
Radun Controls Ltd, White Street, Caerphilly, Glamorganshire

(Illustrated below)

Radun Controls Ltd has introduced two digital voltmeters DM850 and DM851 for general purpose use by engineering laboratories in industrial or educational establishments.

Both instruments have been designed to achieve at a significantly lower cost and without sacrifice to performance or reliability, a four digit voltmeter having an accuracy of 0·1 per cent for d.c. and 0·3 per cent for a.c.

The DM850 universal voltmeter (illustrated) has a coverage of 0 to 1 090V a.c. or d.c. in four ranges, 1·090, 10·90, 109·0 and 1 090V and current 0 to 109mA with full-scale readings of



109·0μA, 1·090, 10·90 and 109·0mA. The sensitivity of both instruments is such that a resolution of 0·02 per cent of full scale is obtained, while the input impedance is 100MΩ on the 1V d.c. range and 30MΩ shunted by 20pF on the 1V a.c. range; for other ranges the input impedance is 2MΩ with the operating frequency range of the instrument 20c/s to 50kc/s. Input signals can be derived from a floating or earthed source and provision is made to protect both types of voltmeter from an inadvertent overload.

Except for the self-balancing servo drive, the voltmeters employ solid state circuits throughout in the form of a 'chopper' drift corrected amplifier, precision voltage reference and a.c. to d.c. convertor.

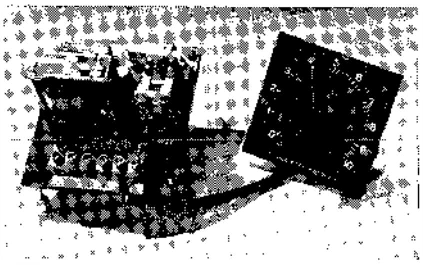
EE 84 759 *for further details*

ELECTRONIC TIMERS

Controls & Automation Ltd, Regal House, 55 Bancroft, Hitchin, Hertfordshire

(Illustrated above right)

These devices, specially designed as alternatives to synchronous motor and trigger type timers, incorporate a unijunction transistor delay circuit operating a heavy duty relay. They can be supplied either chassis only for panel or direct machine mounting, or in a compact sheet steel case.



Time ranges are from 0·05 to 1sec up to 0·25 to 60sec and they are fully adjustable by integral or remote potentiometer. Accuracy is ±1 per cent unaffected by mains variation of ±10 per cent.

They can be arranged to operate the external circuit either during or at the end of the timed period.

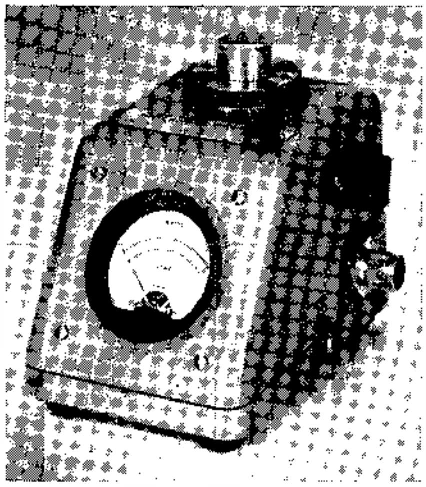
EE 84 760 *for further details*

R.F. POWER MONITOR

Distributed by: Wessex Electronics Ltd, Royal London Buildings, Baldwin Street, Bristol, 1

(Illustrated below)

The new Sierra model 164B bi-directional power monitors give continuous or intermittent indications of incident or reflected power on 50Ω coaxial systems during transmission. Antennas or loads can be matched precisely to transmitters or other power sources simply by monitoring reflected power during operation with the model 164B. Switching from reflected to incident power measurement is simply a matter of turning a knob at the top of the instrument. Four power ranges are available with each insert. Ten plug-in elements allow selection of frequency and power ranges from 2 to 1 000Mc/s and from 100mW to 5kW. Power is read on a direct linear scale, a v.s.w.r. scale is also provided. High directivity and low insertion v.s.w.r. ensure maximum accuracy with minimum disturbance to the transmission line under test. The Sierra movement is damped, no matter what position the insert is in, ensuring ease of transit. Model 164B can be supplied with the new Sierra 'twist off' connectors or to customer requirements.



EE 84 761 *for further details*

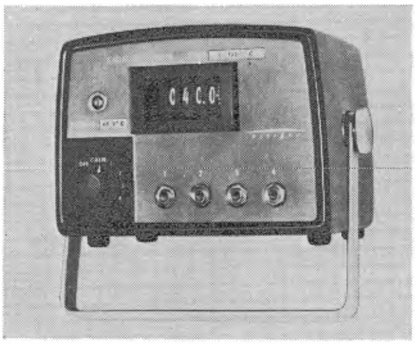
DIGITAL THERMOMETERS

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Illustrated below)

A complete new range of low-cost digital thermometers, giving bi-directional tracking without the use of stepping switches or relays, is now available from Wayne Kerr Laboratories Ltd. High resolution is given by a four-digit illuminated display, with continuous variation indicated on the last figure. Fully interchangeable thermistor probes are available in 17 styles to suit laboratory, medical, meteorological, industrial, biological and chemical applications. Up to four probes can be connected to the instrument at one time, and selection made by a front-panel control.

All thermometers are direct-reading in degrees Fahrenheit or Celsius and can be supplied either for wide-range operation or, for particular applications, with a 5-digit presentation over a specified 20°C or 40°F span. The solid-state circuit can readily be adapted to provide a binary coded decimal output, and a further option is a re-transmitting potentiometer for external scaling.



Styling of the thermometer is in the same compact case (8 × 6 × 6in) as the Digitec voltmeters. Details of platinum resistance thermometers and thermocouple thermometers, in this style, will be released shortly.

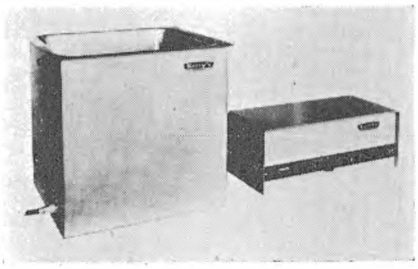
EE 84 762 *for further details*

ULTRASONIC CLEANING EQUIPMENT

Kerry's (Ultrasonics) Ltd, Warton Road, London, E.15

(Illustrated above right)

Kerry's (Ultrasonics) Ltd announce the introduction of their 'Pulsatron' series ultrasonic cleaning equipment, in co-operation with Deltasonics Inc., Hawthorne, California, U.S.A. The ultrasonic generator type KG.850 has a single switch control—high, medium and low—and occupies approximately half a cubic foot. The features of the generator are that the units are fully transistorized and incorporate patented automatic tracking, independent of transducers, and allow the use of high Q, low impedance, sandwich type transducers This arrangement eliminates 'dead spots'—with complete dispersion of energy in the treatment tank. Requiring no warm-up time, it tunes automatically to tank liquid level, cleaning load, or input voltage



and is not affected by open- or short-circuit conditions. All solid state circuits ensure maximum reliability and trouble-free service. The weight of the generator is approximately 35 lb.

The ultrasonic treatment tank is constructed entirely of polished stainless steel material and is complete with drain-off position. The internal dimensions and active area of the tank are 14in by 12in by 12in deep. 'Pulsatron' high efficiency module transducer elements are mounted to the base of the treatment tank and are of the electrostrictive type having sandwich construction of aluminium blocks and lead zirconate titanate ceramic. Nine 'Pulsatron' module transducers are mounted to the base and will withstand an operating liquid temperature of 300°F, which is adequate for all conventional ultrasonic cleaning applications.

This series of ultrasonic cleaning equipment is designed for operation on a frequency of either 25kc/s or 45kc/s, although the initial production units will be offered for operation at 25kc/s. This development will precede the introduction of a 2kW average, 4kW peak ultrasonic cleaning system. Patents applied for.

EE 84 763 *for further details*

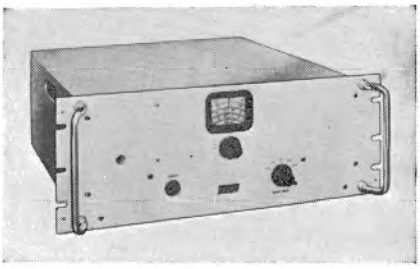
RECEIVER PROTECTION UNIT

Racal Communications Ltd, Western Road, Bracknell, Berkshire

(Illustrated below)

Racal Communications Ltd has recently added to its range of h.f. communication equipment the type MA.197 h.f. receiver pre-selection and protection unit.

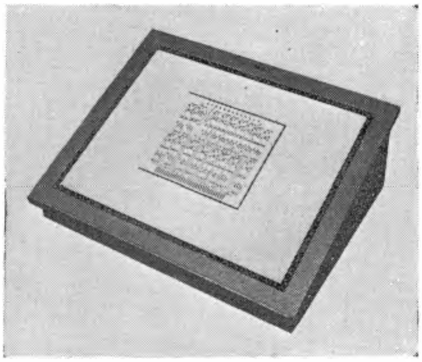
The MA.197 is designed to allow successful operation of receivers in close proximity to high-power transmitters and under similar adverse conditions, where it is essential to reduce breakthrough and prevent physical and electrical damage to the receiver. This condition is unavoidable in ship installation and can also exist where transmitters and receivers must be installed in the same building, or in transportable containers.



When the MA.197 is inserted between the aerial and the receiver input circuits a high degree of pre-selection is achieved, thus reducing adjacent channel interference and cross-modulation products. For an unwanted signal 5 per cent off-tune, an attenuation greater than 95dB is obtained. This is obtained without reduction of overall gain and with only a very small effect upon sensitivity. In addition, the MA.197 protects the receiver against burn-out due to excess r.f. input as it withstands, without damage, radio frequency e.m.f.'s up to 40V.

The MA.197 is tuned locally or remotely by a single knob and covers the range 2 to 24Mc/s with pre-selection and full overload protection, or 1 to 30Mc/s with overload protection only. It has its own power supply suitable for use with all normal a.c. mains, and requires approximately 25W. The unit has a standard 19in front panel for rack mounting, and is suitable for transverse forced-air cooling.

EE 84 764 *for further details*



PRINTED CIRCUIT MASTER GRID

Krisson Printing Ltd, Equipment Division, 184 Acton Lane, London, N.W.10

(Illustrated above)

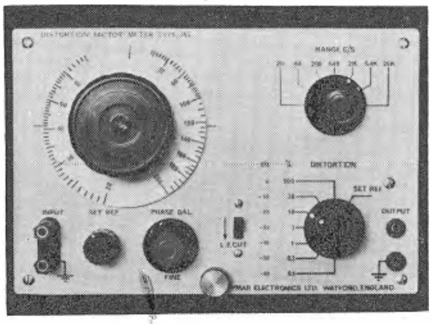
Krisson Printing Ltd, Equipment Division, now offer a range of illuminated tables, incorporating the Krisson master grid, for the preparation of printed circuit masters.

For printed circuits, the master grid is supplied with 0·1in scale, with linear tolerance of ±0·004 over 40in (±0·002in over 20in). These permanent ¼in thick glass grids provide a means of fast and precise drafting. Parallax is eliminated by direct contact between grid and circuit. The illumination from beneath causes the grid lines to show through the circuit sharp and clear for instant guidance.

For instant keying and registering of printed circuits, the master grid can be supplied with the stud and tab register system. The master glass grid has a small stud affixed close to each of the two short edges in the surface of the glass and transparent register tabs are supplied in packets of 50.

Each transparency is fitted loosely over the studs. An accurately punched register tab is then fitted over each stud and taped to the transparency. The operation is then repeated for each successive overlay. Thus the complete set is in perfect alignment with the glass grid lines and each sheet is in register with one another.

EE 84 765 *for further details*



DISTORTION FACTOR METER

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

(Illustrated above)

The latest addition to the Dymar system of plug-in instruments is the type 765 distortion meter. This plug-in unit permits the measurement of the total harmonic distortion of an audio frequency in the range 20c/s to 20kc/s.

The principle employed is that of rejecting the fundamental component of the applied wave and measuring the residue. The voltmeter ranges are calibrated directly in percentage distortion and measurement capability extends from 100 per cent down to 0·025 per cent. Measurements can also be expressed in decibels below fundamental. An l.f. cut is provided to eliminate mains frequency components.

EE 84 766 *for further details*

MINIATURE RELAYS

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustrated above right)

New inexpensive miniature relays, available for a.c. or d.c. and specifically designed to reduce the number of assembled parts have been introduced by Magnetic Devices Ltd. The relays are enclosed in transparent Makrolon covers, which do not support combustion, and mounted directly on the plug base, eliminating the need for independent mounting brackets. Additional support is given by registering the relay body within channels in the cover.

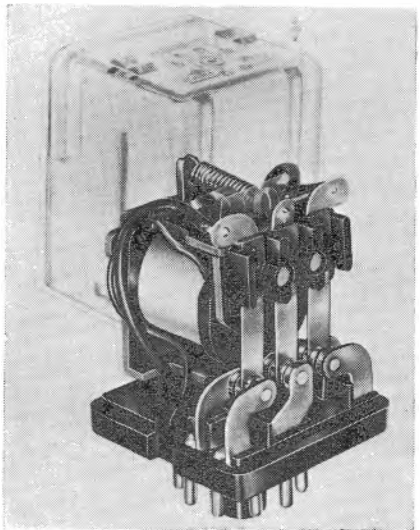
These new relays feature greatly reduced size in height and panel plan, and increased reliability due to the reduction in the number of connexion points. Particular attention has been paid to tracking and air spaces between electrically isolated parts, giving long life when switching arduous loads.

Series 120 (a.c.) and 125 (d.c.) are two pole change-over relays rated at 6A and riveted securely to a moulded 8-pin base. Series 120 has a nominal power of 3VA, maximum working voltage of 250V, and maximum standard coil resistance of 7 300Ω. Series 125 has a nominal power of 1·25W, maximum working voltage of 125V and maximum standard coil resistance of 7 300Ω.

Series 130 (a.c.) and series 135 (d.c.) have three pole changeover contacts as standard, and are mounted on an 11-pin moulded base with adequate anti-tracking barriers. Rated at 6A, series 130 has a nominal power of 3·5VA, maximum working voltage of 250V, a maximum coil resistance of 7 300Ω. Series 135 has a nominal power of 1·5W maximum working voltage of 125V, and a maximum coil resistance of 7 300Ω.

Operate speed of series 125 and 135 is 10msec release 6msec. These new relays will operate in ambient temperature of 40°C maximum, although they can be adapted for high temperature if specified.

Maximum weight is 3 oz. and fine silver or elkonite contacts are available. They can be offered vacuum impregnated for humid or tropical conditions.



EE 84 767 *for further details*

OPTICAL PULSE DIGITIZER

Ferranti Ltd, Thornybank, Dalkeith, Scotland

The Ferranti optical pulse digitizer has been designed to produce electrical pulses whose frequency is proportional to the shaft speed of the driving mechanism, the pulse amplitude being independent of the shaft speed.

The unit consists of a pattern of radial lines on a glass disk which is mounted on the input shaft. Light from a filament lamp is collimated and passed through the disk to a group of photovoltaic cells. A group of photocells is used to minimize the effect of light source variation. Inserted in the light beam is a small piece of glass with a similar pattern of lines as the disk. Moiré fringe patterns are produced by setting the lines on this small piece of glass at an angle to the lines on the input disk. As the disk is rotated these patterns pass over the photocells which produce the required output pulses.

Two, three or four photocells can be provided to allow for direction sensing or frequency doubling and the cell outputs are at approximately 0°, 90°, 180°, and 270° phase relationship to each other. Internal amplifiers can be supplied to amplify these signals if required.

Illumination is provided by a 3V, 3W filament lamp which is under-run at 2·4 to 2·5V to give a life of at least 1 000 hours. The lamp can be replaced without removal of the digitizer.

The present design of the unit gives a total of 100 pulses/revolution/photocell, which may be increased to 200, 400 or 1 000 by electronic circuits.

The development of a range of models is under way and the number of pulses/revolution under consideration is 100, 180, 250, 256, 375, 540, 720, 900, 1 000, 1 080, 1 600, 2 048, 2 250, and 2,700.

In the future lower count models may be added to the range. These would cover 10, 20, 25, 30, 40, 50, and 60 pulses/revolution.

The two main applications of the optical pulse digitizer are as a tachometer—since the output frequency may be read in terms of shaft speed—or to give a measure of shaft angle which can be obtained by counting the pulses from a pre-set datum.

The latter application is suitable for bi-directional counting, such as *X*, *Y* axis positioning on machine tools, the digitizer being driven from a lead screw or any shaft whose angular rotation is proportional to the movement to be controlled or measured.

EE 84 768 *for further details*

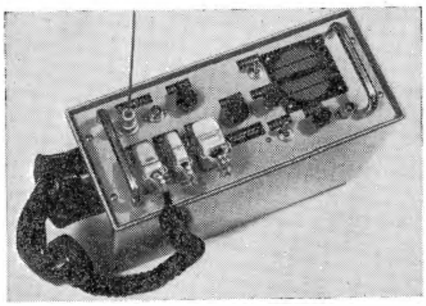
V.H.F. TRANSMITTER-RECEIVER

Radio Communications Co., 16 Abbey Street, Crewherne, Somerset

(Illustrated below)

The 'Telecomm' type TRT/4 v.h.f. radio-telephone is a transportable and completely transistorized instrument for two-way speech communication between similar units or fixed and mobile stations. It is designed for single or double frequency operation on one to six channels in the frequency range 50 to 100Mc/s.

By the use of transistors throughout the receiver and transmitter sections, the power consumption has been reduced to a very low figure in comparison with valve equipment. An internal sealed and



re-chargeable battery pack provides approximately nine hours' use on a 10 to 1 receive to transmit ratio. This battery pack can be quickly removed without opening up the instrument or the use of any tools. If required, this battery can be charged while still in the instrument from external charger type BC/4. Operation can also be carried out from an external 24V d.c. supply such as a heavy duty battery or a.c. mains power unit.

A plug-in ¼ wave whip aerial is used and a coaxial socket is provided for the connexion of an external aerial system or test equipment. A moving-coil speaker is built in.

A plastic case fist microphone with a 'press-to-talk' switch permits simple and speedy operation. When receiving, this microphone also functions as a small speaker. If required a light-weight handset can be supplied in place of the fist microphone. A double headphone assembly with boom microphone can also be supplied.

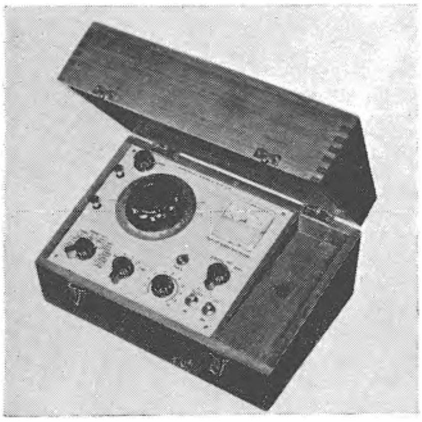
EE 84 769 *for further details*

UNIVERSAL BRIDGE

Thomas Industrial Automation Ltd, Deansgate Lane, Altrincham, Cheshire

(Illustrated below)

This instrument is designed to measure inductance, capacitance and resistance to



an accuracy of ±1 per cent. It is wholly self-contained and transistorized, being powered by internal batteries.

Provision is made to excite the bridge from its internal oscillator at 100c/s, 1kc/s and 10kc/s. For the measurement of resistance of reactive components, for instance, iron cored transformers, the bridge may be excited at d.c., again from internal batteries.

The detector is a moving-coil meter movement, preceded by a high gain amplifier. Automatic gain control is incorporated to facilitate balancing the bridge with inductors and capacitors of low power factor. At d.c. a transistor chopper is switched into the circuit to convert the amplifier to a sensitive d.c. detector

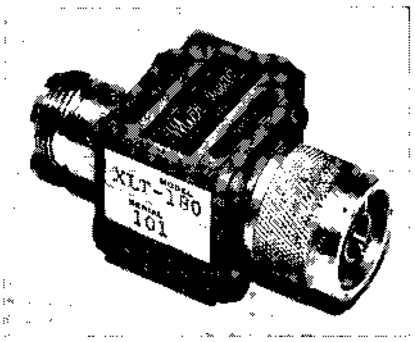
EE 84 770 *for further details*

MINIATURE X-BAND CIRCULATOR

Microwave Systems Ltd, 9-10 River Front, Enfield, Middlesex

(Illustrated above right)

Microwave Systems Ltd has now added a new X-band miniature coaxial circulator/isolator to its already exten-



sive range of Micromega equipment in the U.K.

Operating from 8·5 to 9·5Gc/s with isolation of 25dB, the circulator/isolator has an insertion loss of 0·4dB and v.s.w.r. of 1·15. The component measures 1in × 1¼in × ¾in excluding connectors, and weighs 4 oz including type N connectors.

Similar circulator/isolators are also available for L, S and C bands.

EE 84 771 *for further details*

TRANSISTOR CURVE TRACER

Radar & Communication Instruments Division, Elliott-Automation Ltd, 2-4 Wigton Gardens, Stanmore, Middlesex

(Illustrated below)

The characteristics of a wide range of semiconductor devices can be displayed using the new Elliott transistor curve display, 8100, manufactured by the Radar & Communication Instruments Division of Elliott-Automation.

This compact instrument, which is equally suitable for use in the design laboratory, for production testing or as an educational and training aid, contains a specially designed cathode-ray oscilloscope together with the waveform generators required for testing semiconductor devices under particular conditions.

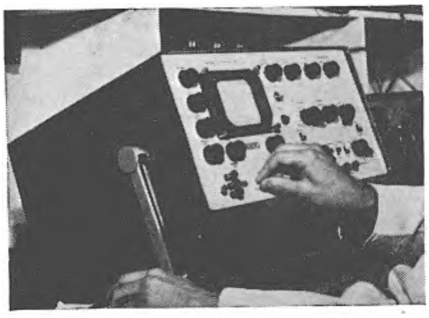
Both npn and pnp transistors may be tested, and similar semiconductors can be compared rapidly to facilitate matching.

Two basic modes of display are available with the new instrument: a staircase waveform input to the device under test will result in the display of a family of curves; or a single step input of continuously variable amplitude can be employed to cover the entire range of currents and voltages, a particularly useful facility when testing thyristors.

The following traces can be produced:

I_c against V_{ce} with I_b or V_{be} as a parameter

I_c against V_{be}



I_c against I_b

I_{ceo} and I_{ces}

and forward and reverse characteristics of Zener diodes, tunnel diodes and rectifiers.

A good deal of information may be gathered from these displays including the avalanche or punch-through voltages of transistors and breakdown characteristics of four-layer diodes and thyristors.

Accurate sources of current and voltage are incorporated to allow the user to calibrate the display before making measurements.

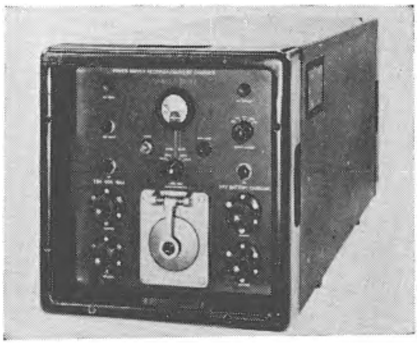
EE 84 772 *for further details*

HEAVY DUTY POWER SUPPLY

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustrated below)

Gresham Lion Electronics Ltd has introduced a heavy duty combined electronic power supply and battery charger. The unit, which is air-cooled, supplies a maximum current of 90A d.c. at a constant voltage of 28V (±1V) from a 200V, 400c/s, three-phase supply. The equipment, known as the GLE 400, has been selected for service by the Ministry of Defence.



The GLE 400 provides constant current for battery charging up to 75A (±10 per cent) in intervals of 15A at voltages varying between 10 and 31V. Ripple is less than 0·5V r.m.s. under all conditions.

The equipment can fulfil both functions at the same time providing that the sum of the currents from the two outputs does not exceed 90A.

The unit, which measures 15½in square by 3ft and weighs 212 lb, is suitable for fixed or vehicular use in temperatures of between −20°C and +50°C in humidities between 0 and 90 per cent.

Input and output fuses and a thermostatic cut-out prevent damage due to overloading or overheating resulting from airflow restriction or air blower failure.

The power supply is housed in a robust aluminium case with recessed control panels and lifting handles. The air inlets are specially shaped and provided with filters to prevent water splash damage.

Maintenance and servicing problems are minimized by mounting all electronic control circuits on three 'plug-in' printed circuit boards.

EE 84 773 *for further details*

Short News Items

The Radar Division of Cossor Electronics Ltd has received an order for the supply of primary radar and associated display and communications system from the Ministry of Aviation. It will provide air traffic surveillance at the Aeroplane and Armament Experimental Establishment at Boscombe Down. As prime contractors Cossor are also responsible for the overall planning and system design including the integration of a standby radar which is being provided by another company and the existing Cossor ACR.6 airfield control radar.

The display facilities to be provided comprise six operational consoles, an equipment room monitor and a radar site monitor. All the displays and associated back-up equipment are taken from the Cossor CRD.100 transistorized display range. Each operational position has inter-console marking facilities and two individual electronic range and bearing markers. Radar information from either the main or standby radars can be selected individually on each display together with appropriate video map information.

The maps to be supplied will be manufactured by the Solartron Electronic Group Ltd. A communications system with controls incorporated in the radar display consoles also forms part of the installation. Remote communications stations are included in addition to the display centre equipment. All communications equipment will be provided by The Telephone Manufacturing Co. Ltd.

The main radar service is to be provided by a Cossor CR.901/D surveillance radar. This is a long range equipment operating at L Band frequencies and is a British version of the Selenia ATCR-2. Two transmitters are employed, normally operating together in frequency diversity, giving an overall mean power of 6kW. Either can be operated on its own to provide a radar service during maintenance periods. Receiving and video processing circuits are also duplicated, and contain the latest techniques in m.t.i., staggered p.r.f., and video integration. The first stage of each receiver is a parametric amplifier, resulting in an overall receiver noise figure of approximately 4dB.

'Radio Frequency Connectors' which is the first part of IEC Publication 169 has recently been published.

Part 1 relates to connectors for r.f. transmission lines for use with electronic equipment and provides general requirements and measuring methods. It establishes uniform requirements for the electrical, climatic and mechanical properties as well as safety aspects; for test methods; for interchangeability and compatibility both between connectors and between connectors and cables; and for classification of connectors into groups according to their ability to withstand extremes of temperature and humidity.

Copies of IEC Publication 169, Part 1, may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price £3 each.

The Twentieth Annual Electronics, Instruments, Controls and Components Exhibition and Convention, organized by the Northern Division of the Institution of Electronics, will be held during the period 28 September to 2 October, inclusive, 1965, at Belle Vue, Manchester. The sessions will be from 10 a.m. to 6 p.m. daily (except for Friday 1 October when the closing time will be 10 p.m.).

The new venue for this Exhibition and Convention has enabled the Exhibition to be considerably enlarged this year with the result that many new companies are able to exhibit for the first time and several former exhibitors have taken the opportunity to rejoin the regular exhibitors. It will also provide adequate car parking facilities away from the traffic-congested centre of Manchester and adjacent to the Exhibition Halls.

Enquiries and applications for tickets should be made to Mr. W. Birtwistle, General Secretary, The Institution of Electronics, 78, Shaw Road, Rochdale, Lancashire.

The 18th Annual Conference and Exhibition on Engineering in Medicine & Biology will be held in Philadelphia, Pa., at the Sheraton Hotel, on 10 to 12 November, 1965.

The meeting is co-sponsored by the Institute of Electrical and Electronics Engineers and the Instrument Society of America, with the participation of the American Society of Mechanical Engineers. The objective of the conference is to contribute to the further advancement of the interdisciplinary area between the engineering, medical and biological sciences, both from an applied and basic standpoint. It will comprise tutorial papers on selected areas of current interest in biomedical engineering, several sessions of contributed papers, informal discussion sessions in the evening and commercial and scientific exhibits.

Further information may be obtained from the conference chairman, Dr. Herman P. Schwan, Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Pa., Exhibitors should contact the conference manager, Mr. F. G. Kettelkamp, Professional Associates, 6520 Clayton Road, St. Louis, Mo., 63117.

The French Government is currently evaluating for use in a number of new aircraft projects a British airborne display system which projects vital flight information directly in the pilot's line of sight.

The system has been supplied by Elliott-Automation and is undergoing flight trials in a Nord-Aviation Atlas at the French Government's Flight Test Centre at Bretigny. Pilots from civil and military establishments in both France and Germany are using the equipment during these trials. A special monitor unit repeats the display seen by the pilot for the benefit of flight trials engineers in the aircraft. The types of symbols used in the display can be changed rapidly to determine the most effective way of presenting information to the pilot during the various stages of flight.

It has also been announced that Sweden has ordered a similar Elliott 'head-up' display system for the System 37 Viggen strike aircraft. It has been stated that over 500 of these aircraft will eventually be in service with the Swedish Air Force.

A LEO 3 computer has been delivered by English Electric-Leo-Marconi Computers Ltd. to the Royal Bank of Scotland's new data processing centre at its head office in St. Andrew Square, Edinburgh. It is the first LEO 3 to be delivered to a bank and the first to be installed in Scotland.

The computer's main tasks will be to maintain current accounts and the register of the bank's stockholders. Lord Nelson of Stafford, chairman and chief executive of English Electric, will formallly open the computer in October.

By the end of the year it is expected that the current accounts at 15 branches will be processed by the LEO 3. The system will be progressively extended throughout the country, the first of the Glasgow branches being scheduled to transfer work to the computer by the first quarter of 1966 and the first of the London branches by the end of the year.

The General Committee of IMEKO held its 7th Session in Warsaw from 9 to 12 June, 1965.

At this session the Member Organizations of IMEKO from Bulgaria, the Chinese People's Republic, Czechoslovakia, the Federal Republic of Germany, the German Democratic Republic, Hungary, Italy, Japan, Poland, Roumania, the Soviet Union, Sweden, the United Kingdom and the United States of America ratified a new Constitution, according to which they reconstituted IMEKO into the International Measurement Confederation.

The current Session of the General Committee also accepted the application of the Egyptian Organization for Standardization representing measurement and instrument specialists in the United Arab Republic, and in this way the first Member Organization from the African continent was included as the 15th Member Organization of the General Committee.

The most important immediate objective of the reconstituted Federation was to outline a structure and a scientific programme of the IV IMEKO Congress to be held in Warsaw in July 1967, in the Palace of Science and Culture.

The Australian Minister for Air has announced completion of a contract to buy two new control and reporting radar systems for the Royal Australian Air Force (RAAF) at a total cost of over £A7 000 000.

The systems will be provided by a group of British and American companies, headed by the British firm of Plessey Radar Ltd as prime contractors and overall managers of the project. The radars will be manufactured by Westinghouse Electric Corporation (US) and the computers by The Marconi Co. Ltd (UK).

The new systems will give the RAAF Defence Organization a greatly enhanced capability to receive early-warning information on air activity, and to carry out surveillance on aircraft movements, the assignment of weapons and the control of interceptions. This increased capability will match the planned operational development of the RAAF.

Incorporation of the most advanced electronic devices will also mean that operational and technical manpower requirements will be considerably less than for the control and reporting units already in operation.

Each of the new control and reporting units will consist of display and data handling equipment, primary three-dimensional and secondary radars, data links, u.h.f., v.h.f. and h.f. communications and power supplies.

The first computer-based alarm analysis scheme in the world is to be installed shortly at the 600MW C.E.G.B. nuclear power station now being built at Oldbury by The Nuclear Power Group. Designed and supplied by AEI Automation Ltd, the scheme is also the first in the world to use cathode-ray tubes for the display of alarm messages. A high-speed AEI digital computer is used as a basis for the system which will be one of the largest on-line computer installations in Great Britain.

In operation, the condition of each of the 3000 plant alarms is checked at least four times every second. Under fault conditions the computer analyses the alarm situation logically and deduces the basic cause alarm. Almost instantaneously, the cathode-ray tubes on the operator's and superintendent's desks display the alarm message together with a statement of the recommended action.

G.E.C. (Telecommunications) Ltd has been awarded a contract by the G.P.O. to supply equipment for a new microwave radio link between London and Leeds. This will add another 1 200 r.f. channel route miles to the existing nationwide network.

Two working r.f. channels and one standby channel will be provided, with automatic changeover from either working channel to the standby in the event of degradation of the signal below a preset level. The equipment will operate in the 6 000Mc/s frequency band and will carry up to 1 800 speech signals on each channel. In addition, the standby channel will be equipped for occasional television service.

The latest G.E.C. 6 000Mc/s microwave radio equipment, in which the r.f. channels are fully semiconductored with the exception of a 10W output travelling wave tube, will be used. All the equipment is battery operated.

Stratton & Company Ltd manufacturers of the Eddystone range of professional radio communication receivers and accessories since 1923, has now been officially renamed Eddystone Radio Ltd. This follows the announcement last March that the Stratton radio interests had been acquired from Laughton and Sons Ltd by English Electric Ltd, and that the company would be operated as a subsidiary of The Marconi Co. Ltd.

A new British Standard has been published giving requirements for epoxide resin bonded glass fibre laminated sheet and phenolic resin bonded paper laminated sheet, which are for use as base material in printed wiring.

The standard (B.S. 3888) specifies preferred nominal thicknesses and tolerances, weights per unit area and conductivities of copper foil, and lays down limits and tests for mechanical and electrical properties. A simple statistical method for defining levels of quality is incorporated in this publication.

Copies of B.S. 3888 'Copper-Clad Synthetic-Resin Bonded Laminated Sheet for Use in Telecommunication and Allied Electronic Equipment' may be obtained from the BSI Sales Branch, 2 Park Street, London W.1. Price 8s. 6d. each. (Postage will be charged extra to non-subscribers.)

The Marconi International Marine Co. Ltd has purchased A.E.I.'s marine radio business. Marconi Marine will continue to market A.E.I. marine radio equipment for as long as may be necessary and, with the agreement of individual owners, will also assume responsibility for A.E.I.'s rental-maintenance and maintenance contracts for the servicing and operation of existing ship installations, including those with radar, hitherto handled by A.E.I.

Facilities for data transmissions between Britain and the USA over the public telephone network are now included in the Post Office Datel 600 service, introduced earlier this year. Negotiations to extend international data transmission facilities are now in progress with a number of European administrations and agreements are expected to be concluded in the near future.

Data signalling at the rate of 600 bits/sec should be possible on all international Datel 600 calls, rising to a maximum of 1200 bits/sec in many cases.

To set up an international Datel 600 call, one subscriber will be able to dial over ISD or be connected through appropriate overseas telephone exchange.

At the INEL 65, the second International Exhibition of Industrial Electronics, to be held from 7 to 11 Septembed 1965 in the halls of the Swiss Industries Fair in Basle, the extensive display of exhibits will be international in scope: more than 450 exhibitors have announced their intention of taking part; they will come from 13 different countries and will display the products of more than 600 manufacturers.

The BBC has placed a contract with British Insulated Callender's Construction Co. Ltd for the design, supply and erection of a 1 000ft mast to carry the aerials at the u.h.f. television transmitting station for BBC-2 which is to be built near Waltham-on-the-Wolds, Leicestershire. This station, hitherto identified in preliminary announcements as the Nottinghamshire station and which the BBC hopes to bring into operation during the winter of 1966/67, will extend the BBC-2 service to much of Nottinghamshire and Leicestershire and parts of Lincolnshire and Derbyshire.

Crime detection and prevention experiments carried out by Liverpool City Police using closed circuit television have been so successful that they have placed an order for a further four television channels with EMI Electronics Ltd.

The cameras will be used by the Police Commando Force to observe places where crime is likely to occur, such as unattended car parks and storage areas.

As they are fitted with zoom lenses, the cameras can be located some distance from the areas being observed. Once set up they require little attention and their outline can be disguised to blend with the local architecture.

Close to each camera site is a temporary control room in which a detective watches the television pictures produced by the cameras. If a suspect is picked up on the screen the detective sends information by radio to Police Headquarters, or to the nearest member of the Commando Force in the area.

A Symposium on Microminiaturization in Automatic Control Equipment and in Digital Computers will be held in Munich from 21 to 23 October 1965. The Symposium will be organized by the VDI/VDE-Fach-gruppe Regelungstechnik in conjunction with the Nachrichtentechnische Gesellschaft im VDE and sponsored jointly by the International Federation of Automatic Control (IFAC) and the International Federation for Information Processing (IFIP). Apart from some survey papers, more then 50 papers received from Great Britain, USA, the Netherlands, USSR, CSSR, Switzerland and Germany will be presented and discussed. The programme is classified under the following headings: Pneumatic and Hydraulic Components, Optical Components, Storage Components, Integrated Circuits, Construction and Interconnexion Techniques, Computer Techniques, Measuring Devices and Transducers. All participants will receive preprints of the papers a few weeks prior to the Symposium.

Enquiries should be addressed to the Verein Deutscher Ingenieure, Abt. Organization, 4 Dusseldorf 10, Postfach 10250.

Details of a new award for the Invention of the Year have been released by the Institute of Patentees and Inventors.

The best invention submitted will receive the Richardson Gold Medal under a Trust provided by Mr. A. W. Richardson, Chairman of the Council of the Institute.

Successful inventors may also become eligible for scholarships, apprenticeships and financial assistance under a separate Trust which is under consideration.

Closing date for entries for the 1965 Invention Award is 31 December, and full conditions are obtainable from the Institute of Patentees and Inventors, Abbey House, Victoria Street, S.W.1.

A postgraduate course in quantum electronics leading to an M.Sc. or diploma (D.U.S.) will commence in October 1965. The course will be concerned with the mechanism of operation, and the design and application of quantum electronic devices such as masers and lasers and will consist of lectures, research seminars, laboratory work and a research project. The diploma will be awarded on the results of an examination held in the following June. University graduates may be permitted to continue with a research project for a further three months and submit a dissertation for an M.Sc.

In addition to the above course the existing post-graduate course covering the broad field of electronics will continue to be held.

Further particulars of both courses, and of research opportunities in the Department of Electronics, may be obtained from the Academic Registrar, The University of Southampton.

The London Electrical Manufacturing Co. Ltd which currently manufactures a large range of capacitator types, has recently completed a study of thin film silicon monoxide capacitors. This study was undertaken on their behalf by Cambridge Consultants Ltd, a company which offers a liaison service between industry and research workers in the University.

A complete thin film plant was designed by the advisors of Cambridge Consultants Ltd and manufactured by their workshop staff. After eight months' operation a team of two men have achieved satisfactory and reproducible results. The plant is one of the largest operating in England, having a bell-jar of more than 1000 litres capacity and enabling deposition, with mask-changing, over an area of approximately one square foot.

During the study a variety of peripheral equipment was developed, including a new continuous feed aluminium evaporation source, and a direct reading crystal ratemeter. The aluminium source employs a specially developed vacuum motor driving a capstan that feeds aluminium wire on to a hot filament. Evaporation rate is continuously variable up to 4g/sec.

When the study was commissioned, towards the end of 1963, it appeared that thin film deposition might rapidly become a very attractive manufacturing technique for capacitors. Experience has now shown that conventional, hydrid and integrated circuit techniques are to be preferred, and the manufacturing plant is soon to be available for other thin film studies.

The 'Early Bird' satellite was inaugurated as part of a commercial telephone system on 28 June by a telephone conversation between President Johnson, in Washington, and the Prime Minister in London.

Also linked by telephone in a big international ceremony were the Prime Minister of France (M. Georges Pompidou), the German Chancellor (Dr. Ludwig Erhard), the President of Switzerland (Professor Hans-Peter Tschudi) and the Italian Minister of Posts and Communications (Signor Carlo Russo).

The M.E.L. Equipment Co. Ltd is manufacturing a sample number of radiosondes to a new Meteorological Office design.

M.E.L. have also received from the Meteorological Office a contract for a data logger which will initially be used in the evaluation of these prototype equipments.

The logger will record on punched paper tape temperature, humidity and atmospheric pressure data transmitted simultaneously from two balloon-borne radiosondes. It will also include a channel for recording radar tracking data for the calculation of balloon position. An internal channel for recording timing pulses will also be provided.

A research contract worth £95 000, the first to be placed with a university by the Ministry of Technology, has been awarded to the University of Newcastle-upon-Tyne for a three- to four-year programme of work on the application of electronic digital computers to type composition and printing. Exploratory work carried out at Newcastle University during the past three years under grant aid from the former Department of Scientific and Industrial Research will be extended to cover actual industrial applications.

One of the objects of the work is to provide the printing industry with a series of adaptable computer programmes which perform all the operations of copy preparation, editing and actual type composition. An equally important aim is to provide a wholly automatic system of preparing material from computers for publication in an improved, readable form without running the risk of introducing errors by human data handling.

'Packaging of Covered Winding Wires for Electrical Purposes', B.S. 1489, has been extensively revised. Since the issue of the 1952 edition there have been important developments in this sphere of packaging—notably the growing trend towards the adoption of bulk packages in the form of cylindrical containers ('paks'), and long traverse reels for end pay-off ('Long Traverse Vertical'—LTV—reels). The new edition of B.S. 1489 deals with both these forms of packaging, together with reels for all covered winding wires. It specifies dimensions, materials, construction and maximum eccentricity (where appropriate) for each type, and includes requirements for the marking of reels and the filling, packing and labelling of reels and containers. The standard also makes recommendations on the appropriate reels and containers for different wire sizes.

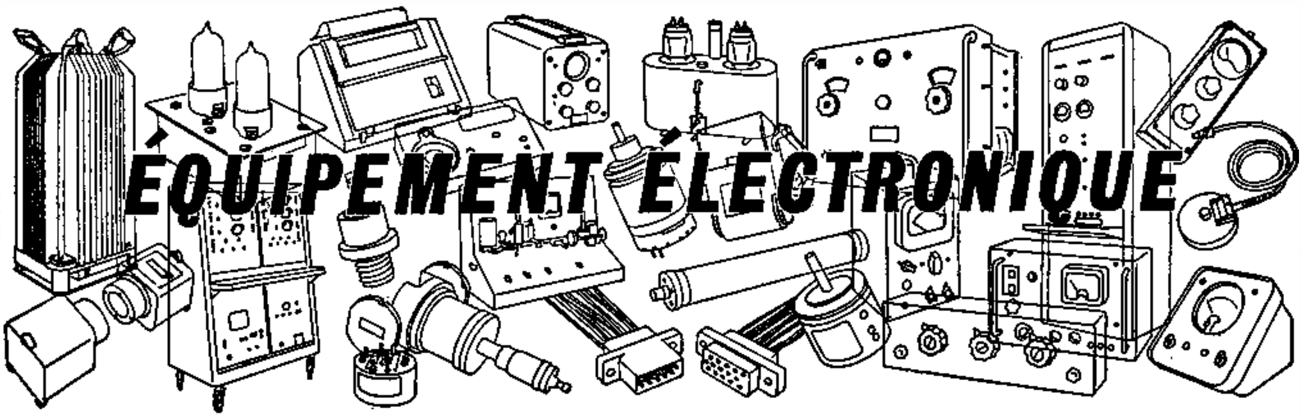
B.S. 1489 forms one of a series of standards dealing with reels and drums for various types of wire.

Copies of B.S. 1489 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 6/- each. (Postage will be charged extra to non-subscribers).

The Society of Cardiological Technicians of Great Britain is holding its seventeenth Exhibition of Cardio-Pulmonary Apparatus at the Londoner Hotel, Welbeck Street, London, W.1, from 4.0 to 9.0 p.m. on Friday 5 November and from 9.30 a.m. to 1.0 p.m. on Saturday 6 November, 1965.

Approximately twenty exhibitors will be showing the latest equipment in this specialized field.

Tickets can be obtained from Margaret Hole, Exhibition Secretary, Cadiac Research Department, Guy's Hospital, London, S.E.1.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 550 à 555

ANALYSEUR DE COUSSINETS

Cosmocord Ltd, Eleanor Cross Road, Waltham Cross, Hertfordshire

(*Illustration à la page 550*)

Cet analyseur a été étudié pour permettre aux ingénieurs d'entretien de matériel d'usine de déterminer rapidement l'état comparatif des coussinets dans les machines-outils. L'analyseur indique également le degré de surcharge lorsqu'un outil de coupage avance trop rapidement dans le matériau traité. Il obvie à la nécessité de devoir changer des coussinets à des intervalles prédéterminés et empêche ainsi la production de pièces usinées défectueuses.

Il comprend un dispositif sensible habituellement monté sur le logement des coussinets de la machine et un dispositif de lecture.

Pour le contrôle de production automatique, on a prévu un dispositif indicateur muni d'un disjoncteur de surcharge pouvant être utilisé pour arrêter le processus en cas de défaut ou donner une alarme appropriée. L'ensemble du système est de construction robuste et, dans des conditions d'emploi normales, il est pratiquement indestructible, le dispositif sensible étant unpetit élément solide et le circuit de lecture étant entièrement transistorisé.

EE 84 751 *pour plus amples renseignements*

VOLTMÈTRE NUMÉRIQUE

Hewlett Packard, Dallas Road, Bedford

(*Illustration à la page 550*)

Ce nouveau voltmètre numérique entièrement constitué de corps solides, modèle 3460A, offre une combinaison de grande précision, à immunité de bruit, à impédance d'entrée constante et élevée; et une grande vitesse, par l'emploi d'un nouveau principe de fonctionnement.

La lecture du voltmètre est à cinq chiffres, un sixième chiffre servant à l'indication hors-gamme. La résolution maxima sur la gamma la plus basse est de 10 μV et l'impédance d'entrée est constante à 10 MΩ sur toutes les gammes. La précision totale ($\pm$0,005 % de la lecture, $\pm$2 comptages, dans des environments de +10 à +40° C) est maintenue, même à la plus grande vitesse de fonctionnement de l'appareil, jusqu'à 15 lectures par seconde.

Le modèle 3460A est un appareil à la fois intégrateur et potentiométrique. C'est, dans un sens, un voltmètre automatique, intégrateur et différentiel. Il effectue la lecture en deux périodes d'échantillonnage rapides et successives. Dans la première période, un convertisseur flottant et gardé voltage-fréquence intègre l'inconnue, produisant une série d'impulsions dont le taux est proportionnel à la tension d'entrée instantanée. Ces impulsions sont comptées et le comptage met en oeuvre les chiffres les plus significatifs de la lecture. C'est alors que le comptage total est emmagasiné dans les éléments de comptage puis transféré à un convertisseur numérique-analogique. Ce dernier produit un signal analogique exactement égal au comptage emmagasiné (lecture) et ce signal est comparé à l'inconnue. La différence est transmise au convertisseur tension-fréquence et elle est intégrée en impulsions qui sont comptées durant la deuxième période d'échantillonnage. Ces comptages sont enregistrés dans le cinquième et le sixième chiffres. Le comptage total est alors transféré à l'affichage.

L'instrument est potentiométrique à tous égards car il produit un signal proportionnel qu'il compare avec l'entrée, s'appuyant principalement pour la précision sur des rapports de résistance de précision et sur une tension stable de référence. Contrairement à d'autres instruments de mesure potentiométriques, il est, cependant, relativement exempt des bruits de mode normal, sans qu'il fasse usage de filtres d'entrée réducteurs de vitesse. Le rejet de mode commun, à déséquilibre de 1 kΩ, est de 140 dB à toutes les fréquences.

N'importe laquelle des quatre gammes, ou gammes automatiques, peut être choisie par bouton poussoir. Etant donné que le modèle 3460A comporte l'indication hors-gamme à 20 % sur toutes les gammes, il peut mesurer jusqu'à 1200 V c.c., la tension la plus élevée étant présentée avec une entière précision sur l'affichage à six chiffres. Des compteurs réversibles sont utilisés pour l'intégration des signaux variant autour de zéro.

Le fonctionnement des systèmes a été facilité par l'inclusion d'une sortie d'imprimeur à chiffres codés binaires (1-2-2-4), la programmation à distance de la gamme et de la mesure, et des accessoires pour la mesure de la résistance et du courant alternatif. Les commutateurs de panneau frontal transfèrent le fonctionnement aux bornes arrière.

EE 84 752 *pour plus amples renseignements*

ÉLÉMENTS LOGIQUES À LAME VIBRANTE

Flight Refuelling Ltd, Wimborne, Dorset

(*Illustration à la page 550*)

Pour répondre au besoin de systèmes de commutation à logique simple, combinant l'entrée et la sortie multiples avec une vitesse de commutation assez rapide et un prix modéré, la société Flight Refuelling Ltd a mis au point une gamme d'éléments logiques à lame vibrante, les "RELOG", basés sur le commutateur miniature à lame vibrante Hamlin MRC-1.

Pour certains types d'applications, le système logique à lame vibrante peut être utilisé avec profit, particulièrement dans l'industrie des machines-outils et de manutention mécanique, car bien que sa vitesse de fonctionnement soit inférieure à la logique transistorisée, dans de nombreuses applications dans ces domaines ce sont les dispositifs d'entrée et de sortie qui constituent en fait les facteurs limitatifs de vitesse pour n'importe quel système complet.

Le système logique à lame vibrante constitue donc un système à vitesse suffisante et à bas prix, si l'on tient compte du fait que l'équipement auxiliaire nécessaire est moins compliqué et qu'une seule ligne de puissance est exigée. Une compacité approchant celle des dispositifs logiques transistorisés peut être réalisée car un maximum de 36 éléments Relog peuvent être montés sur plaquette de 25,4 cm × 25,4 cm.

Deux fonctions logiques, "OU" et "NI", sont normalement prévues, toutes deux sous forme d'éléments à deux ou à quatre entrées. Les deux fonctions présentent, en outre, l'avantage de pouvoir augmenter indéfiniment le nombre d'entrées en ajoutant des diodes supplémentaires.

Grâce à l'emploi d'un commutateur à lame vibrante pouvant recevoir des courants atteignant 0,5 A, un grand nombre de charges logiques (25 au maximum) peuvent être actionnées à partir d'un élément et l'isolement électrique complet du circuit de sortie de chaque élément est ainsi réalisé.

EE 84 753 *pour plus amples renseignements*

ANALYSEUR À 100 CANAUX

Distributeurs: Nuclear Enterprises (G.B.) Ltd, Sighthill, Edimbourg, Ecosse

(Illustration à la page 551)

L'analyseur à 100 canaux Spectrascope Modèle 100, est le "dernier-né" de la série des systèmes d'analyse spectrométrique nucléaire Laben et il est maintenant livré par la société Nuclear Enterprises (G.B.) Ltd.

Cet appareil portatif, de grande précision et à usages multiples, se compose d'un amplificateur, d'un convertisseur, d'une mémoire logique, d'un dispositif de décodage pour l'impression et d'une échelle de comptage automatique. Il effectue des additions aussi bien que des soustractions et les enregistreurs de contenu sont accessibles en série ou, sur demande, en parallèle. Il peut être utilisé dans les modes suivants: opérations non-aléatoires, intégration multiple, échantilonnage, calcul des moyennes et analyse des hauteurs d'impulsions. Il peut, en outre, être utilisé comme élément à canal unique dans le mode à échelle multiple et il comporte un amplificateur incorporé RC à écrêtage et à disjoncteur à maxima. Il peut être mis en action avec un imprimeur en série et, sur demande, avec un imprimeur en parallèle. La spécification prévoit une capacité de comptage de 10^6-1, un taux de calcul numérique de 10 MHz/sec, une durée de commutation de 1,5 μsec, un cycle d'emmagasinage de 6 μsec et un temps mort total maximum de 17 μsec.

Un élément à fiches, Modèle 100A, pouvant être ajouté à l'analyseur de base ou utilisé indépendamment, est également livrable. L'élément 100A comprend une alimentation en haute tension pour détecteurs nucléaires, un programmateur-minuterie pour la présélection de la mesure et les temps d'attente, un élément de sélection de canaux pour l'intégration des données et un sélecteur de signaux de sortie pour la stabilisation automatique en fonction de l'élément extérieur.

EE 84 754 *pour plus amples renseignements*

VOLTMÈTRE NUMÉRIQUE

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 551)

La société Dawe Instruments Limited fournit maintenant un nouveau convertisseur c.c.-c.a. et un voltmètre numérique à bas prix. Ces deux appareils, type 653 et 652 respectivement, peuvent être utilisés ensemble ou séparément pour fournir une lecture à quatre chiffres de tension continue ou alternative d'une précision de 0.1 ou de 0,2 %.

Le voltmètre numérique type 652 est offert en deux versions de base, soit le type 652A et le type 652AX. La première emploie une simple tension de référence à cellule standard pour mesurer jusqu'à 0,2 mV, avec une précision totale de ±0,2 % de lecture maxima sur chaque gamme. Le type 652AX utilise une référence à diode Zéner pour mesurer jusqu'à 0,1 mV avec une précision totale de ±0,1 % de la lecture maxima. Les deux versions comprennent quatre grammes donnant des lectures maxima de 1, 10, 100 ou 1000 V c.c. sur compteurs mécaniques à indication décimale automatique. L'impédance d'entrée est de 2,2 MΩ, ce qui est suffisant pour éviter de surcharger la plupart des circuits.

La conception de ces appareils s'inspire des systèmes d'asservissement à auto-équilibrage, la tension continue d'entrée étant injectée à un amplificateur comparateur, de même qu'une tension continue de référence contrôlée par un moteur entraîné par potentiomètre. L'amplificateur comparateur produit une sortie de tension alternative dont l'amplitude et la phase correspondent à la différence entre les niveaux de signal et de référence, cette sortie étant injectée au bobinage de commande d'un moteur électrique biphasé. Le moteur entraîne les compteurs d'affichage et déplace le curseur du potentiomètre jusqu'à ce que la tension de référence adapte l'entrée de tension continue au comparateur.

Le temps de lecture est en moyenne de 2 sec, avec un maximum de 5 sec (pour lecture maxima). L'instrument comporte un dispositif de calage du zéro. Il peut être actionné par alimentation secteur de 50 ou 60 Hz et il est enfermé dans un coffret métallique bleu portatif, suffisamment solide pour pouvoir être employé en atelier. Il mesure 19,05 cm × 15,24 cm × 27,94 cm de haut. Son poids est d'environ 4 kg.

Le convertisseur c.a.-c.c., type 653, permet la mesure précise de tensions alternatives au moyen du voltmètre numérique type 652. Son circuit se compose d'un atténuateur d'entrée, d'un amplificateur à gain élevé et d'un étage de sortie à filtre de moyenne. Les signaux d'entrée peuvent être appliqués au convertisseur à partir de niveaux de millivolts jusqu'à 500 V efficaces, à des fréquences de 50 Hz à 25 kHz. L'entrée peut être flottante, fournissant un maximum de 1 V c.c. sur les trois gammes inférieures et 0,5 V c.c. sur la gamme supérieure. La précision nominale est de ±0,3 % de la sortie maxima sur chacune des gammes de 50 Hz à 10 kHz et, à ±0,4 % près, jusqu'à 25 kHz. La linéarité sur toutes les gammes est de 0,05 %. L'appareil fonctionne sur courant secteur et il est logé dans un coffret assorti au coffret de l'appareil type 652.

EE 84 755 *pour plus amples renseignements*

CONTRÔLEUR DE CÂBLES TÉLÉPHONIQUES

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustration à la page 551)

Ce contrôleur de câbles téléphoniques, autonome et portatif, offre des avantages appréciables aux ingénieurs d'installations de câbles. Il a été conçu par la société Standard Telephones and Cables Ltd.

Portant la référence 74226, cet appareil peut mesurer rapidement la capacitance mutuelle, le déséquilibre de capacitance et le déséquilibre de pourcentage de résistance. On peut donc mesurer les diverses combinaisons de déséquilibre de capacitance dans des sections de câbles à construction en paires ou en étoile, de sorte que des configurations optimum à jonction de section peuvent être employées en vue de minimiser la diaphonie ou les interférences.

Les gammes de mesure sont les suivantes:

Déséquilibre de capacitance—

Gamme 1—280-0-280 pF
Gamme 2—1100-0-1100 pF

les deux gammes ayant une précision de ±(1 % + 2 pF). Une capacitance mutuelle allant jusqu'à 0,112 μF peut être mesurée avec une précision de ±50 pF. Pour un déséquilibre de résistance de ±0,5 % du circuit conducteur, la résistance peut être mesurée avec une précision de 0,01 %.

Les caractéristiques de fonctionnement comportent un dispositif permettant de faire basculer la boîte pour obtenir une meilleure visibilité et l'éclairage facultatif de l'échelle, pour pouvoir l'employer sous une mauvaise lumière. En outre, la sortie à fréquence acoustique dans les écouteurs est limitée lorsque le pont n'est pas en équilibre, de manière à éviter l'inconfort de fonctionnement.

Le circuit en pont fonctionne en liaison avec un oscillateur et un amplificateur transistorisés. Il existe deux versions ayant des fréquences de fonctionnement de 800 et 1000 Hz respectivement. La puissance est fournie intérieurement par deux batteries sèches de 4,5 V.

L'appareil est logé dans un coffret en bois muni de poignées et d'un couvercle démontable. Il mesure 43,18 cm × 30,48 cm × 25,4 cm et son poids est de 19,5 kg.

EE 84 756 *pour plus amples renseignements*

OSCILLOSCOPE

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(Illustration à la page 552)

Le nouvel oscilloscope RO 501 à tube cathodique de 127 mm fonctionne sur 3,3 kV stabilisés pour assurer des images d'une grande brillance à n'importe quelle vitesse de balayage. Il ne comporte qu'une seule plaquette de circuit imprimé dont tous les composants sont codés.

Sa largeur de bande s'étend du courant continu à 6 MHz, son temps de montée est de 55 nsec, sa sensibilité de 50 mV/cm à 20 V/cm, couverte en neuf gammes étalonnées (1, 2, 5 degrés), et sa précision de mesure est de ±5 %.

La gamme de balayage va de 1 μsec/cm à 100 msec/cm en six degrés étalonnés et une commande à variation continue de 10:1 étend la gamme jusqu'à 1 sec/cm; un dispositif de grossissement variable offre une vitesse de balayage maxima de 200 nsec/cm. Un circuit de déclenchement d'une grande souplesse d'emploi permet de choisir le point de déclenchement, cependant qu'un intégrateur d'images de télévision permet de déclencher la base de temps à partir d'impulsions d'images de télévision. La sensibilité interne est de 5 mm et la sensibilité externe est de 4 V de crête à crête.

L'appareil est muni d'un étalonneur intérieur, le micromètre à deux couleurs est éclairé sur les côtés et l'oscilloscope peut recevoir des caméras normales sur entr'axes de 12,7 cm. Les dimensions hors-tout de l'oscilloscope sont: 25,08 cm de hauteur × 37,46 cm de profondeur × 19,05 cm de largeur. Son poids est de 8 kg environ.

EE 84 757 *pour plus amples renseignements*

COUPEUR ET DÉNUDEUR DE FILS MÉTALLIQUES

Multicore Solders Ltd, Multicore Works, Maylands Avenue, Hemel Hempstead, Hertfordshire

(Illustration à la page 552)

Ce nouvel appareil à couper et à dénuder les fils métalliques, modèle "Bib," est muni d'une jauge présélectrice pour le choix de n'importe quelle épaisseur de fil de 12 à 26. En faisant un usage correct du sélecteur, l'isolement peut être dénudé sans danger d'endommagement au fil. Cet outil comprend des poignées plastiques capitonnées pouvant être facilement empoignées.

EE 84 758 *pour plus amples renseignements*

VOLTMÈTRES NUMÉRIQUES

Radun Controls Ltd, White Street, Caerphilly, Glamorganshire

(Illustration à la page 552)

La société Radun Controls Ltd a créé deux voltmètres numériques, DM850 et DM851, pour l'emploi universel par les laboratoires de construction dans les établissements industriels ou d'enseignement.

Les deux instruments ont été conçus pour offrir, à un prix nettement inférieur et sans sacrifier le rendement ou la fiabilité, un voltmètre à quatre chiffres dont la précision est de 0,1 % pour le courant continu et de 0,3 % pour le courant alternatif.

Le voltmètre universel DM850 (voir notre gravure) couvre 0 à 1090 V c.a. ou c.c. en quatre gammes, 1,090, 10,90, 109,0 et 1090 V et un courant de 0 à 109 mA, avec des lectures sur la totalité de l'échelle de 109,0 μA, 1,090, 10,90 et 109,0 mA. La sensibilité des deux instruments est telle qu'une résolution de 0,02 % de l'échelle totale est obtenue, tandis que l'impédance d'entrée est de 100 MΩ sur la gamme de 1 V c.c. et de 30 MΩ shuntés par 20 pF sur la gamme de 1V c.a. Pour d'autres gammes, l'impédance d'entrée est de 2 MΩ avec une gamme de fréquence de fonctionnement de 20 Hz à 50 kHz. Les signaux d'entrée peuvent être obtenus à partir d'une source flottante ou à la masse et les deux modèles de voltmètre peuvent être protégés contre toute surcharge accidentelle.

A l'exception du mécanisme d'entraînement asservi, à auto-équilibrage, les voltmètres emploient des circuits constitués de corps solides sous forme d'amplificateur à correction de dérive et à relais modulateur, d'une référence de tension de précision et d'un convertisseur c.a.-c.c.

EE 84 759 *pour plus amples renseignements*

MINUTERIES ÉLECTRONIQUES

Controls & Automation Ltd, Regal House, 55 Bancroft, Hitchin, Hertfordshire

(Illustration à la page 552)

Ces instruments, spécialement conçus pour servir de variantes aux minuteries à moteur synchrone et du type à déclenchement, comportent une circuit à retard à transistors à unijonction faisant oeuvre de relais de puissance. Ils peuvent être fournis soit pour montage direct sur panneau ou sur machine, soit dans un coffret compact en acier.

Les gammes de temps vont de 0,05 à 1 sec et de 0,25 à 60 sec et elles sont entièrement règlables au moyen d'un potentiomètre intégral ou à distance. La précision est de ± 1 % et elle est insensible aux variations de courant secteur ± 10 %.

Ces minuteries peuvent actionner le circuit externe, soit pendant la période minutée soit à la fin de cette période.

EE 84 760 *pour plus amples renseignements*

CONTRÔLEUR DE PUISSANCE HF

Distributeurs: Wessex Electronics Ltd, Royal London Buildings, Baldwin Street, Bristol, 1

(Illustration à la page 552)

Les nouveaux contrôleurs de puissance bidirectionnels Sierra, modèle 164B, donnent une indication continue ou intermittente de la puissance incidente ou réfléchie sur 50 systèmes coaxiaux durant la transmission. Les antennes ou les charges peuvent être adaptées avec précision aux transmetteurs ou à d'autres sources de puissance en contrôlant simplement la puissance réfléchie durant le fonctionnement avec le modèle 164B. La commutation de la mesure de puissance réfléchie à la mesure de puissance incidente s'effectue simplement en tournant un bouton sur le haut de l'instrument. Quatre gammes de puissance sont prévues. Dix éléments à fiches permettent le choix de la fréquence voulue et des gammes de puissance de 2 à 1000 MHz et de 100 mW à 5000 W. La puissance est indiquée sur une échelle linéaire directe; une échelle de rapport d'amplitude de tension est également prévue. La haute directivité et le rapport d'amplitude de tension à faible insertion assurent le maximum de précision avec le minimum de dérangement de la ligne de transmission soumise au contrôle. Le mouvement Sierra est attenué, quelle que soit la position de l'insertion, ce qui facilite le transit. Le modèle 164B peut être fourni avec les nouveaux connecteurs Sierra ou selon la spécification du client.

EE 84 761 *pour plus amples renseignements*

THERMOMÈTRES NUMÉRIQUES

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 553)

Une gamme complète de nouveaux thermomètres numériques à bas prix, assurant le pointage bidirectionnel sans l'emploi de relais à degrés ou de commutateurs, est fournie maintenant par The Wayne Kerr Laboratories Ltd. Une haute résolution est garantie par l'affichage éclairé à quatre chiffres, avec variation continue indiquée sur le dernier chiffre. Des sondes à thermistor pleinement interchangeables sont prévues en 17 modèles pour les applications médicales, de laboratoire, météorologiques, industrielles, biologiques et chimiques. Jusqu'à quatre sondes peuvent être reliées à l'instrument en même temps et la sélection s'effectue par une commande de panneau frontal.

Tous les thermomètres sont à lecture directe en degrés Fahrenheit ou Centigrade et ils peuvent être fournis soit pour le fonctionnement à gamme étendue, pour des applications particulières, soit avec présentation à cinq chiffres. Le circuit constitué de corps solides peut être aisément adapté pour fournir une sortie décimale codée binaire sur 20° C ou 40° F. On peut, en outre, prévoir un potentiomètre de retransmission pour la démultiplication extérieure.

Le thermomètre est dans le même boîtier compact (20,32 cm × 15,24 cm × 15,24 cm) que les voltmètres Digitec. Des détails sur les thermomètres à résistance de platine et les thermométres à thermocouples, du même genre, seront données prochainement.

EE 84 762 *pour plus amples renseignements*

ÉQUIPMENT DE NETTOYAGE À ULTRASONS

Kerry's (Ultrasonics) Ltd, Warton Road, London, E.15

(Illustration à la page 553)

La société Kerry's (Ultrasonics) Ltd vient d'annoncer l'introduction de son matériel de nettoyage à ultrasons de la série "Pulsatron". Ce matériel est introduit en collaboration avec la Deltasonics Inc., Hawthorne, Californie, U.S.A. Le générateur à ultrasons type KG.850 comporte une commande à un seul commutateur—haut, moyen et bas—et il occupe un volume d'environ 14dm³. Les éléments du générateur sont entièrement transistorisés et à règlage automatique breveté, indépendant des transducteurs, ce qui permet l'emploi de transducteurs du type sandwich, à faible impédance et à facteur de qualité élevé. Cet arrangement élimine les "points morts", avec dispersions complète de l'énergie dans le réservoir de traitement. N'ayant besoin d'aucun temps de chauffe, l'appareil s'accorde automatiquement au niveau du liquide du réservoir, de la charge de nettoyage ou de la tension d'entrée, et il est insensible aux conditions de circuit fermé ou ouvert. Les circuits entièrement constitués de corps solides assurent le maximum de fiabilité et de fonctionnement sans le moindre ennui. Le poids du générateur est d'environ 17 kg.

Le réservoir de traitement par ultrasons est entièrement en acier inoxydable et il est complet avec point de vidange. Les dimensions internes et la zone active du réservoir sont de 35,56 cm × 30,48 cm × 30,48 cm. Les éléments à transducteurs modulaires de grande efficacité "Pulsatron" sont montés sur la base du réservoir de traitement et ils sont du type électrostrictif, étant construits sous forme de "sandwichs" de blocs d'aluminium et de céramique de titanate de zirconate de plomb. Neuf transducteurs modulaires "Pulsatron" sont montés à la base et ils peuvent résister à une température de liquide de fonctionnement de 300°F, suffisante pour toutes les applications de nettoyage à ultrasons de type classique.

Ce matériel de nettoyage à ultrasons a été étudié pour pouvoir fonctionner à une fréquence de 25 kHz ou de 45 kHz, selon le choix, mais les éléments de production initiale ont été prévus pour l'utilisation à 25 kHz. Ils précèdent l'introduction d'un système de nettoyage à ultrasons à 2 kW de moyenne et 4 kW de pointe, pour lequels un brevet a été demandé.

EE 84 763 *pour plus amples renseignements*

ELÉMENT DE PROTECTION POUR RÉCEPTEUR

Racal Communications Ltd, Western Road, Bracknell, Berkshire

(Illustration à la page 553)

La société Racal Communications Ltd vient d'ajouter à sa gamme de matériel de communications HF l'élément de protection et de présélection pour récepteur HF type MA 197.

Le MA.197 a été conçu pour pouvoir utiliser au mieux les récepteurs se trouvant à proximité étroite d'émetteurs de grande puissance ou, dans des situations défavorables analogues, lorsqu'il est nécessaire de réduire les interférences et d'empêcher l'endommagement physique et électrique du récepteur. Cet état de choses est inévitable dans les installations de navires et peut se produire également lorsqu'il faut installer des émetteurs et des récepteurs dans le même bâtiment ou dans des conteneurs transportables.

Lorsque le MA.197 se trouve placé entre l'antenne et les circuits d'entrée du récepteur, on obtient un degré élevé de présélection car on réduit ainsi les interférences du canal adjacent et la transmodulation. Pour un signal indésirable de 5 % hors-accord, on obtient une atténuation supérieure à 95 dB, sans réduire le gain total et avec un très léger effet sur la sensibilité. De plus, le MA.197 protège le récepteur contre l'épuisement dû à l'entrée HF excessive, car il peut résister, sans dommage, à des force électromagnétiques haute fréquence allant jusqu'à 40 V.

Le MA.197 est accordé localement ou à distance par un seul bouton et il couvre la gamme de 2 à 24 MHz avec présélection et disjonction à maxima, ou de 1 à 30 MHz avec disjonction seule. Il a son propre bloc d'alimentation convenant pour toutes les tensions alternatives secteur normales; sa consommation électrique est d'environ 25 W. L'appareil comporte un panneau frontal standard de 48,26 cm pour montage sur bâti et il peut être utilisé pour le refroidissement transversal par air forcé.

EE 84 764 *pour plus amples renseignements*

GRILLE PRINCIPALE À CIRCUIT IMPRIMÉ

Krisson Printing Ltd, Equipment Division, 184 Acton Lane, London, N.W.10

(Illustration à la page 553)

La société Krisson Printing Ltd, Equipment Division, offre maintenant une gamme de tables, comprenant la grille principale Krisson, pour la préparation de matrices de circuits imprimés.

Pour les circuits imprimés, la grille principale est fournie avec échelle de 0,1 pouce, avec tolérance linéaire de ±0,004 pouce sur 40 pouces (±0,002 pouces sur 20 pouces). Ces grilles en verre de 6 mm d'épaisseur constituent une matrice sûre et précise. Les parallaxes sont éliminés par le contact direct entre la grille et le circuit. L'éclairage par en-dessous permet de voir les lignes de la grille à travers le circuit de manière claire et distincte pour le guidage instantané.

Pour la manipulation et l'enregistrement instantanés des circuits imprimés, la grille principale peut être fournie avec le système à registre à tenon et oreille. La grille de verre principale comprend ainsi une petit tenon fixé à proximité de chacune des deux petits bords de la surface du verre; des tenons transparents sont livrables en paquets de 50 pièces.

Les transparences sont montées de manière lâche sur les tenons. Une orelle perforée avec précision est alors fixée sur chacun des tenons et rattachée à la transparence. On répète cette opération pour chacune des appliques. Ainsi l'ensemble complet est aligné de manière parfaite aux lignes de la grille de verre et chacune des feuilles est "en registre" avec l'autre.

EE 84 765 *pour plus amples renseignements*

INSTRUMENT DE MESURE DU FACTEUR DE DISTORSION

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

(Illustration à la page 554)

Le distorsiomètre type 765 constitue la plus récente réalisation dans la gamme des instruments à fiches du système Dymar. Il permet de mesurer la distorsion totale en harmoniques d'une fréquence acoustique dans la gamme de 20 Hz à 20 kHz.

Le principe utilisé consiste à rejeter la composante fondamentale de l'onde appliquée et à mesurer le résidu. Les gammes voltmétriques sont étalonnées directement en pourcentages de distorsion et la capacité de mesure s'étend de 100 % à 0,025 %. La mesure peut être également exprimée en décibels au-dessous de la valeur fondamentale. Une coupure de basse fréquence est prévue pour éliminer les composantes de fréquence secteur.

EE 84 766 *pour plus amples renseignements*

RELAIS MINIATURE

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(Illustration à la page 554)

Des relais miniature à bas prix, prévus pour tension alternative ou continue, et spécifiquement conçus pour réduire le nombre de pièces assemblées, ont été réalisés par la société Magnetic Devices Ltd. Ces relais sont enfermés dans des housses transparentes en Makrolon qui ne supportent pas la combustion. Ils sont montés directement sur la base à fiches, obviant ainsi à la nécessité d'avoir des supports de montage indépendants. Un support supplémentaire est assuré en fixant le corps du relais dans des canaux à l'intérieur de la housse.

Ces nouveaux relais se caractérisent par leur encombrement fort réduit et leur fiabilité accrue due à la réduction du nombre de points de connexion. Un soin particulier a été apporté aux prises de courant et aux entrefers entre parties électriquement isolées, ce qui assure une longue durée de vie lors de la commutation de charges élevées.

Les relais de la série 120 (c.a.) et

125 (c.c.) sont des composants bipolaires à permutation et à puissance nominale de 6 A. Ils sont solidement rivetés à une base moulée à huit broches. La série 120 a une puissance nominale de 3 VA, une tension maxima de régime de 250 V et une résistance de bobine standard maxima de 7300 Ω. La série 125 a une puissance nominale de 1,25 W, une tension maxima de régime de 125 V et une résistance de bobine standard maxima de 7300 Ω.

Les relais de la série 130 (c.a.) et 135 (c.c.) comportent des contacts tripolaires normaux à permutation, et ils sont montés sur une base moulée à 11 broches avec barrières appropriées de prises de courant. Quant aux relais de la série 130, leur tension de régime est de 250 V, leur puissance nominale de 3,5 VA et leur résistance de bobine maxima de 7300 Ω.

La vitesse de fonctionnement des séries 125 et 135 est de 10 msec avec déclenchement de 6 msec. Ces nouveaux relais fonctionnent dans des températures ambiantes de 40° C au maximum, bien qu'ils puissent être utilisés à des températures plus élevées, sur demande.

Leur poids maximum est d'environ 94 grammes et ils sont munis de contacts en elkonite ou en argent fin. Il sont livrables avec imprégnation sous vide pour des atmosphères humides ou tropicales.

EE 84 767 *pour plus amples renseignements*

ENREGISTREUR NUMÉRIQUE D'IMPULSIONS OPTIQUES

Ferranti Ltd, Thornybank, Dalkeith, Scotland

L'enregistreur numérique d'impulsions optiques Ferranti a été réalisé pour produire des impulsions électriques dont la fréquence est proportionnelle à la vitesse d'arbre du mécanisme d'entraînement. L'amplitude des impulsions est indépendante de la vitesse d'arbre.

L'appareil comporte un réseau de lignes radiales sur un disque de verre monté sur l'arbre d'entrée. La lumière d'une lampe à filament est collimatée et transmise à travers le disque à un groupe de cellules photo-voltaïques. Ce groupe est utilisé pour minimiser l'effet des variations de sources de lumière. Dans le faisceau lumineux se trouve une petite pièce de verre comportant un réseau de lignes semblables à celles du disque. Des réseau marginaux sont produits en fixant les lignes de cette petite pièce de verre à un certain angle par rapport aux lignes du disque d'entrée. A mesure que le disque tourne, ces réseaux passent au-dessus des cellules photoélectriques qui produisent les impulsions de sortie voulues.

Deux, trois ou quatre cellules photoélectriques peuvent être fournies pour la perception de la direction ou le doublage de la fréquence et les sorties de cellules sont à environ 0°, 90°, 180° et 270° de phase les unes des autres. Des amplificateurs internes peuvent être fournis pour amplifier ces signaux, si nécessaire.

L'éclairage est assuré par une lampe à filaments de 3 V, 3 W, utilisée à une puissance inférieure, soit 2,4 à 2,5 V, pour lui assurer une vie d'au moins 1 000 heures. L'ampoule peut être remplacée sans qu'il soit nécessaire d'enlever l'enregistreur numérique.

Le modèle actuel de l'appareil fournit un total de 100 impulsions/tour/cellule photoélectrique, pouvant être augmentées à 200, 400 ou 1 000 impulsions par des circuits électroniques.

La mise au point d'une série de modèles est en cours et l'on envisage de porter le nombre des impulsions par tour aux chiffres suivants: 100, 180, 250, 256, 375, 540, 720, 900, 1 000, 1 080, 1 600, 2 048, 2 250 et 2 700.

Des modèles à comptage inférieur pourront être ajoutés par la suite. Ces derniers effectueraient 10, 20, 25, 30, 40, 50 et 60 impulsions par tour.

Les deux principales applications de l'enregistreur d'impulsions optiques sont celles de tachymètre—étant donné que la fréquence de sortie peut être lue en tant que vitesse d'arbre—et de mesureur de l'angle d'arbre. Cette dernière fonction s'effectue en comptant les impulsions à partir d'une donnée préréglée. Elle est particulièrement indiquée pour le comptage bi-directionnel, tel que le positionnement d'axe X, Y sur machines-outils, l'enregistreur numérique étant actionné par une vis de plomb ou n'importe quel arbre dont la rotation angulaire est proportionnelle au mouvement devant être contrôlé ou mesuré.

EE 84 768 *pour plus amples renseignements*

ÉMETTEUR-RÉCEPTEUR POUR HYPER-FRÉQUENCES

Radio Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Illustration à la page 554)

Le radio-téléphone VHF "Telecomm." type TRT/4, est un instrument portatif et entièrement transistorisé de communication à deux directions entre éléments semblables ou entre stations fixes ou mobiles. Il a été étudié pour le fonctionnement à fréquence double ou simple sur une ou six voies dans la gamme de fréquences de 50 à 100 MHz.

Grâce à l'emploi de transistors dans les sections d'émission ou de réception, la consommation électrique a été réduite à un niveau extrêmement bas pas rapport au matériel à lampes. Un bloc à batteries interne, scellé et rechargeable, assure environ 9 heures d'utilisation sur un rapport réception-émission de 10 à 1. Ce bloc à batterie peut être facilement retiré sans ouvrir l'instrument ou employer des outils. Si nécessaire, cette batterie peut être chargée pendant qu'elle est encore dans l'instrument, à partir d'un chargeur extérieur type BC4. L'appareil peut aussi être mis en action à partir d'une alimentation extérieure de 24 V c.c., telle que celle fournie par une batterie de puissance ou un bloc d'alimentation secteur en courant alternatif.

L'antenne fouet à fiches d'un quart d'onde et la douille coaxiale dont est muni l'appareil permettent de le relier à un système à antenne extérieure ou à un élément de contrôle. Il comprend également un haut-parleur à cadre mobile incorporé.

Un microphone de format réduit à coffret en matière plastique avec commutateur à pression permet l'utilisation simple et aisée du "Telecomm" type TRT/4. Durant la réception, ce microphone fonctionne également comme petit haut-parleur. On peut fournir, sur demande, un combiné à main à la place du petit microphone. Un assembage à du petit microphone. Un assemblage à perche peut aussi être fourni.

EE 84 769 *pour plus amples renseignements*

PONT UNIVERSEL

Thomas Industrial Automation Ltd, Deansgate Lane, Altrincham, Cheshire

(Illustration à la page 555)

Cet instrument a été étudié pour mesurer l'inductance, la capacitance et la résistance à un degré de précision de ±1%. Il est entièrement autonome et transistorisé et entraîné par des batteries internes. Il peut être excité par son oscillateur interne à 100 Hz, 1 kHz et 10 kHz. Pour mesurer la résistance de composants réactifs comme, par exemple, les transformateur à noyau de fer, le pont peut être excité au courant continu, une fois encore par les batteries internes.

Le détecteur est un instrument de mesure à cadre mobile, précédé d'un amplificateur à gain élevé. La commande de gain automatique est incorporée à l'appareil afin de faciliter l'équilibrage du pont avec des inducteurs et des condensateurs à faible facteur de puissance. Au courant continu, un relais modulateur est branché au circuit pour convertir l'amplificateur en un détecteur de courant continu sensible.

EE 84 770 *pour plus amples renseignements*

CIRCULATEUR MINIATURE À BANDE X

Microwave Systems Ltd, 9-10 River Front, Enfield, Middlesex

(Illustration à la page 555)

La société Microwave Systems Limited vient d'ajouter un nouveau circulateur-isolateur coaxial miniature à bande X à sa gamme déjà fort étendue de matériels Micromega dans le Royaume-Uni.

Fonctionnant entre 8,5 et 9,5 GHz, avec isolement de 25 dB, le circulateur-isolateur a une perte par insertion de 0,4dB et un rapport d'amplitude de tension de 1,15. Il mesure 2,54 cm × 2,85 cm × 1,90 cm sans les connecteurs et pèse 124,5 grammes avec les connecteurs de type N.

Des circulateurs-isolateurs analogues sont également livrables pour les bandes L, S et C.

EE 84 771 *pour plus amples renseignements*

TRACEUR DE COURBES DE TRANSISTORS

Radar & Communication Instruments Division, Elliott-Automation Ltd, 2-4 Wigton Gardens, Stanmore, Middlesex

(Illustration à la page 555)

Les caractéristiques d'une gamme étendue de semi-conducteurs peuvent être affichées à l'aide du nouveau traceur de courbes de transistors, Elliott 8 100, fabriqué par la Radar & Communication Instruments Division de la société Elliott-Automation Ltd.

Cet instrument compact, pouvant être utilisé tout aussi bien en laboratoire que pour le contrôle de production ou comme instrument d'enseignement, contient un oscilloscope à rayons cathodiques de conception spéciale ainsi que des générateurs de formes d'ondes pour le contrôle des semi-conducteurs dans des conditions particulières.

Il peut contrôler tant des transistors npn que pnp, et des semiconducteurs analogues peuvent être comparés rapidement pour faciliter l'équilibrage.

Le nouvel instrument offre deux modes d'affichage de base; une entrée de forme d'onde en escalier reliée au dispositif soumis à l'essai produira l'image d'un groupe de courbes; ou alors une entrée à un degré d'une amplitude à variation continue pourra être employée pour couvrir la gamme entière de courants et de tensions, ce qui constitue un avantage particulièrement apprécié lors du contrôle des thyristors.

Les traces suivantes peuvent être produites:

I_c contre V_{ce} avec I_b ou V_{be} comme paramètre
I_c contre V_{be}
I_c contre I_b
I_{ceo} et I_{ces}

ainsi que les caractéristiques d'avance et d'inversion des diodes Zéner, des diodes tunnel et des redresseurs.

Un grand nombre de renseignements peuvent être tirés des ces instruments, dont les tensions d'avalanche et de perforation de transistors et les caractéristiques de panne de diodes à quatre couches et de thyristors.

Des sources précises de courant et de tension sont également prévues pour permettre à l'utilisateur d'étalonner l'appareil avant d'effectuer des mesures.

EE 84 772 *pour plus amples renseignements*

BLOC D'ALIMENTATION DE PUISSANCE

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustration à la page 555)

La société Gresham Lion Electronics a réalisé un bloc d'alimentation et un chargeur de batterie combinés de puissance. Cet ensemble type GLE 400, à refroidissement par air fournit un courant continu maximum de 90 A à une tension constante de 28 V (±1 V), à partir d'une alimentation triphasée de 200 V, 400 Hz. Il a été adopté par le Ministère de la défense britannique pour ses services.

Le GLE 400 fournit un courant constant pour batterie, chargeant jusqu'à 75 A (±10%), à intervalles de 15 A et à des tensions variant entre 10 et 31 V. Le facteur d'ondulation est inférieur à 0,5 V efficace dans toutes les conditions.

L'équipement peut remplir les deux fonctions en même temps, à condition que la somme des courants des deux sorties ne dépasse pas 90 A. Il mesure 49,37 cm sur les côtés et 91,4 cm de hauteur; son poids est de 97 kg. Il peut être employé de manière fixe ou véhiculaire dans des températures allant de −20°C à +50°C et dans une humidité de 0 à 90%.

Les fusibles d'entrée et de sortie et le coupe-circuit thermostatique empêchent l'endommagement dû à la surcharge ou à l'échauffement excessif, résultant d'une limitation du flux d'air ou d'une panne du ventilateur.

Le bloc d'alimentation est logé dans un robuste coffret en aluminium avec panneaux de contrôle en retrait et poignées de levage. Les orifices d'air sont de forme spéciale et munis de filtres pour empêcher tout dégât dû aux éclaboussures d'eau.

Les problèmes d'entretien et de dépannage sont réduits au minimum en raison du fait que tous les circuits de contrôle électronique sont montés sur trois plaquettes de circuit imprimé à fiches.

EE 84 773 *pour plus amples renseignements*

Résumés des Principaux Articles

Un amplificateur à volume constant et à gamme dynamique étendue par A. G. Morris

Résumé de l'article aux pages 502 à 507

Les progrés réalisés dans la conception d'amplificateurs à volume constant on été limités par la nécessité de mettre au point un dispositif idéal à gain variable. Le transistor saturé se rapproche étroitement du modèle idéal, car il ne produit qu'une très faible distorsion et sa gamme de contrôle potentiel dépasse 120dB. L'auteur décrit un système utilisant un transistor saturé qui comprime une variation d'entrée de 60dB dans 4dB. L'emploi de cette gamme dynamique étendue présente cependant deux problèmes supplémentaires: en premier lieu, il faut établir les conditions nécessaires à l'obtention d'une marge de stabilité optima d'un système de contrôle à gamme étendue et, en second lieu, il faut des circuits supplémentaires pour empêcher l'amplification du bruit entre les paroles. Des solutions à ces deux problèmes sont indiquées; la solution du premier de ces problèmes étant d'usage général dans la théorie du contrôle.

Oscillateurs à diodes tunnel microondes par J. Swift

Résumé de l'article aux pages 508 à 511

L'auteur traite de l'utilisation des diodes tunnel dans les oscillateurs microondes. Il examine, notamment, les paramètres des diode tunnel ainsi que leur emploi dans les circuits d'oscillateurs microondes. Après avoir donné des indications de réalisation détaillées pour un oscillateur de 2 GHz, il fournit une liste bibliographique d'ouvrages parus à ce sujet.

Dispositifs à plusieurs ouvertures pour les applications logiques entièrement magnétiques par D. J. Morris

Résumé de l'article aux pages 512 à 519

Cet article décrit le comportement d'un certain nombre de dispositifs à plusieurs ouvertures de forme différente pour les opérations logiques à combinaisons de base. Ces dispositifs peuvent être combinés de manière à leur permettre de réaliser des opérations logiques en séquence sans nécessiter de composants supplémentaires autres que le fil métallique servant à relier les dispositifs. Le système utilisé pour le couplage des dispositifs porte le nom de "MAD-R". Il autorise des tolérances opérationnelles étendues et il a été étudié en vue d'applications exigeant une grande fiabilité.

Le comportement et les limitations des amplificateurs à niveau réduit employant des diodes tunnel par B. Easter

Résumé de l'article aux pages 520 à 523

L'amplificateur a diode tunnel trouve une utilisation potentielle comme simple étage d'entrée de récepteur pour les systèmes au-dessus de 1 GHz où un facteur de bruit modérément réduit est acceptable. Le comportement de ce type d'amplificateur est étudié dans cet article en fonction des besoins des systèmes de relais-radio à ligne de vision. L'auteur examine, entre autres, les effets non-linéaires pouvant se produire, ainsi que la conversion m.a./m.p.

Il étudie les problèmes que pose la réalisation d'un amplificateur pratique, ainsi que celui de la mise au point d'un circuit stable. Il donne, enfin, quelques détails d'un amplificateur prévu pour 2 GHz.

Un circuit de mesure simple du gain de courant à grand signal de transistors et à présentation numérique par S. L. Hurst

Résumé de l'article aux pages 524 à 528

L'auteur décrit un circuit d'essai simple à courant continu qui permet l'affichage direct, à l'aide de n'importe quel fréquencemètre numérique ordinaire, de la valeur du gain de courant à grand signal d'émetteur commun de transistors à divers niveaux de courant. La précision de la mesure est comparable à celle des contrôleurs de transistors à courant continu du type commercial normal, mais sa lecture est plus claire et facile. On peut aussi mesurer les courants de fuites d'émetteurs communs et l'équilibrage du gain de courant de transistors peut être effectué aisément avec une grande précision d'adaptation.

Contrôle automatique du gain de bruit par J. S. Bowman et H. Sutcliffe

Résumé de l'article aux pages 529 à 531

L'application du contrôle de gain automatique aux amplificateurs de bruit est compliquée par les caractéristiques inappropriées des détecteurs à retard de prédétection. Les auteurs analysent la réponse de ces détecteurs lorsque le bruit est normalement réparti et ils concluent que l'amplification de post-détection et un retard sont nécessaires pour le contrôle de gain automatique efficace. Ils décrivent le circuit de contrôle d'un amplificateur générateur de bruit dans la zone de 10-100kHz. L'élément à gain variable est un diviseur de potentiel à diode avec polarisation de courant contrôlée.

Un amplificateur paramétrique dégénéré à refroidissement par azote de 6 GHz pour usage radiométrique
par D. Chakraborty et G. F. D. Millward

Résumé de l'article aux pages 532 à 535

Il s'agit ici d'un amplificateur paramétrique dégénéré de 6 GHz à refroidissement par azote liquide conçu pour un radiomètre en vue de la mesure, au moyen d'une "étoile-radio", du gain de l'antenne d'une station de sol d'un satellite de communication. Les deux étages, dont le gain est de 15 dB et dont la largeur de bande est de 200 MHz par rapport aux points de -3 dB, sont en cascade afin d'assurer un gain minimum de 30 dB. La largeur de bande inhérente de l'amplificateur s'est avérée suffisante pour l'utilisation prévue bien que des circuits spéciaux furent nécessaires pour étendre la largeur de bande. La température de souffle de l'amplificateur refroidi est de 57°K.

La limite théorique de sensibilité d'un système radiométrique utilisant cet amplificateur est brièvement analysée.

Système d'échantillonnage pour la stabilisation à un degré élevé des moteurs à courant continu par F. Szlavik et G. Orbán

Résumé de l'article aux pages 536 à 539

Ce système a été réalisé pour la mise en marche automatique et la stabilisation de vitesse des moteurs à courant continu de moyenne et de faible puissance. Les circuits électroniques décrits en détail ont été étudiés pour un moteur entraînant un relais interrupteur à faisceau de neutrons. La précision et la stabilité de fréquence et de phase sont déterminées dans des gammes étendues de paramètres influents par la précision et la stabilité d'un générateur de signaux de référence.

La distorsion et la sortie de puissance d'amplificateurs à modulation de durée d'impulsions par E. C. Bell et T. Sergent

Résumé de l'article aux pages 540 à 542

Les méthodes de modulation de largeur d'impulsions sont examinées du point de vue de leur application aux amplificateurs acoustiques linéaires à haute efficacité. Le spectre du signal modulé est analysé et les résultats que l'on obtient indiquent le degré de distorsion auquel on peut s'attendre et la manière de réduire cette distorsion par un choix approprié de fréquence d'étalonnage.

Il est démontré qu'avec une largeur de bande acoustique de 15KHz et une modulation à un seul bord, une fréquence d'échantillonnage non-inférieure à 90KHz est nécessaire si l'on veut que la sortie ne contienne pas de composants d'amplitude supérieurs à 1% du signal acoustique. Avec la modulation à double bord, la fréquence d'échantillonnage se réduit à 75KHz.

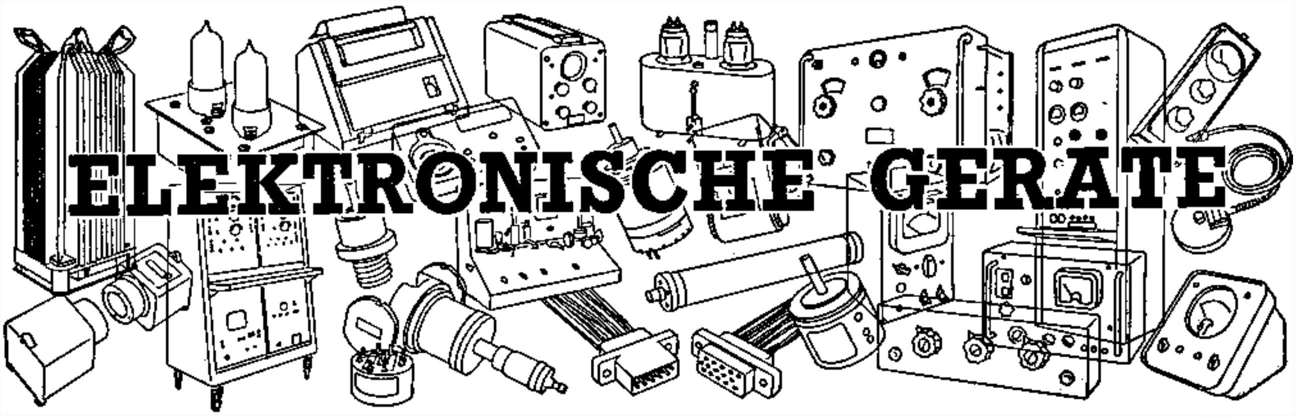
Il est également démontré qu'en pratique une sortie de puissance de l'ordre de 20W serait le maximum pouvant être économiquement obtenu avec des transistors ordinaires.

Un instrument d'affichage d'intervalles de temps réciproques à circuits transistorisés par R. H. Kay

Résumé de l'article aux pages 543 à 545

L'instrument en question a été conçu pour les applications neurophysiologiques et il peut enregistrer la fréquence de répétition "instantanée" ainsi que chacune des impulsions qui se produisent. Le débit est affiché sur un tube à électrons-volts et une indication linéaire du taux de répétition est fournie de 7 à 7000 sec^{-1}.

ELEKTRONISCHE GERÄTE



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 550 bis 555

Lageranalysator

Cosmocord Ltd, Eleanor Cross Road, Waltham Cross, Hertfordshire

(Abbildung Seite 550)

Diese Instrumentserie wurde entwickelt, um dem Betriebsingenieur zu ermöglichen, schnell den verhältnismässigen Zustand von Lagern usw. in Werkzeugmaschinen zu bestimmen. Das Gerät zeigt auch die Überlastung von Schneidwerkzeugen an, die zu schnell in den zu bearbeitenden Werkstoff vordringen. Durch diese Analysatoren wird die Notwendigkeit, Lager in vorbestimmten Zeitintervallen auszuwechseln, beseitigt, und sie schützen gleichzeitig gegen Fertigung fehlerhaft bearbeiteter Teile.

Das Gerät besteht aus einem Tastkopf, der üblicherweise direkt auf das Lagergehäuse montiert wird, und einer Anzeigeeinrichtung.

Für automatische Fertigungsregelung gibt es auch eine Anzeigeeinrichtung mit angebautem Überlastungsschützen, der den Arbeitsablauf bei Auftreten eines Fehlers stoppt oder eine geeignete Warnung gibt. Das ganze System ist robust gebaut und unter normalen Bedingungen praktisch unzerstörbar, da der Tastkopf eine kleine massive Einheit und die Anzeigeeinrichtung volltransistorisiert ist.

EE 84 751 *für weitere Einzelheiten*

Digitalvoltmeter

Hewlett Packard, Dallas Road, Bedford

(Abbildung Seite 550)

Das durchweg in Festkörpertechnik ausgeführte neue Digitalvoltmeter 3460A bietet durch Anwendung eines neuen Arbeitsprinzips eine Kombination von grosser Genauigkeit, Rauschfreiheit, hoher und konstanter Eingangsimpedanz und grosser Messhäufigkeit.

Das Voltmeter hat eine volle 5stellige Anzeige mit einer sechsten Stelle für Überschreiten des Grenzwertes. Die maximale Auflösung ist im niedrigsten Bereich 10 μV, die Eingangsimpedanz in allen Bereichen konstant 10 MΩ. Die volle Messgenauigkeit des Gerätes ($\pm$0,005 % der Anzeige $\pm$2 Signale bei +10°...+40°C Umgebungstemperatur) wird selbst bei der grössten Messhäufigkeit von bis zu 15 Messungen je Sekunde aufrechterhalten.

Das Modell 3460A ist sowohl integrierend wie kompensierend. Es ist in gewisser Hinsicht ein automatisierter, integrierender Differenzspannungsmesser mit Schutzschaltung, der Messungen in zwei schnellen, aufeinanderfolgenden Perioden vornimmt. In der ersten Periode integriert ein massefreier Spannungs-Frequenzumsetzer die Unbekannte und erzeugt eine Impulsreihe, deren Frequenz der momentanen Eingangsspannung proportional ist. Die Impulse werden gezählt und bestimmen die bedeutsamsten Ziffern der Anzeige. Das Gesamtergebnis der Zählung wird gespeichert und das gespeicherte Ergebnis in einen Digital-Analogwandler übertragen, der ein der gespeicherten Zählung genau äquivalentes Analogsignal abgibt, das mit der Unbekannten verglichen wird. Die Differenz wird in den Spannungs-Frequenzumsetzer gespeist; die abgegebenen integrierten Impulse werden in der zweiten Periode gezählt, und das Ergebnis geht in die 5. und 6. Stelle des Endergebnisses ein. Das komplette Ergebnis der Zählung wird nunmehr in die Anzeigeeinheit übertragen.

Das Gerät ist in jeder Hinsicht selbstkompensierend, erzeugt ein proportionales Signal, das es mit dem Eingangssignal vergleicht, und stützt sich zur Erzielung seiner Genauigkeit auf Präzisionswiderstandsverhältnisse und eine konstante Bezugsspannung. Ungleich anderen Kompensationsmessgeräten ist es jedoch verhältnismässig unempfindlich gegen Rauschen in der normalen Betriebsweise, ohne dass man auf Eingangsfilter, die die Geschwindigkeit herabsetzen, zurückgreifen muss. Gleichtaktunterdrückung ist bei allen Frequenzen und 1 kΩ Unsymmetrie 140 dB.

Jeder der vier Teilbereiche oder Bereichautomatik kann mittels Drucktaste eingestellt werden. Da beim Modell 3460A der Bereichendwert in allen Bereichen um 20 Prozent überschritten werden kann, ist die messbare, auf der 6stelligen Anzeige mit voller Genauigkeit dargestellte Höchstspannung 1200 V—. Zur Integration von Signalen, die um Null schwanken, verwendet man umkehrbare Zähler.

Verwendung in Systemen wird durch den vorhandenen binärkodierten Ausgang (1-2-2-4) für einen Zifferndrucker erleichtert; Bereich und Messfunktion lassen sich fernprogrammieren, und es gibt Zubehör für das Messen von Wechselspannungen und Widerständen. Schalter auf der Frontplatte übertragen Bedienung auf die auf der Rückseite liegenden Anschlüsse.

EE 84 752 *für weitere Einzelheiten*

Zungenschalterlogik

Flight Refuelling Ltd, Wimborne, Dorset

(Abbildung Seite 550)

Um die Nachfrage nach einem einfachen logischen Schaltsystem mit Mehrfacheingängen und -ausgängen mit angemessen schneller Schaltzeit bei mässigem Preisaufwand zu decken, hat Flight Refuelling Ltd die auf dem Hamlin-Miniaturzungenschalter MRC-1 beruhenden logischen Zungenschalterelemente "Relog" entwickelt.

In gewissen Industriezweigen, hauptsächlich Werkzeugmaschinenbau und Fördertechnik, ist die Anwendung der Zungenschalterlogik vorteilhaft, trotzdem sie langsamer arbeitet als die transistorisierte Logik. Für viele Anwendungsmöglichkeiten auf diesen Gebieten wird die Schaltgeschwindigkeit jedes beliebigen kompletten Systems durch die Ein- und Ausgabeeinrichtungen beschränkt.

Die Zungenschalterlogik gibt daher bei geringerem Aufwand ein ausreichend schnelles System, dessen erforderliche Hilfseinrichtungen weniger kompliziert sind und das nur eine Speiseleitung benötigt. Die erreichbare Packdichte entspricht näherungsweise der der transistorisierten Logik, da bis zu 36 Relog-Bausteine auf eine 254 × 254-mm-Karte montiert werden können.

Zur Zeit sind zwei logische Funktionen lieferbar: "ODER" und "NOR" in Ausführungen mit zwei und vier Eingängen. Durch Nachrüstung weiterer Dioden lässt sich bei beiden Funktionen die Anzahl der Eingänge unbegrenzt erhöhen.

Die Verwendung von Zungenschaltern, durch für Ströme bis zu 0,5 A bemessen sind, kann eine grosse Anzahl logischer Lasten (bis zu 25 maximal) von einem Element getrieben werden. Die Ausgangsschaltung jedes Elementes ist elektrisch vollkommen isoliert.

EE 84 753 *für weitere Einzelheiten*

Analysator für 100 Kanäle

Vertrieb: Nuclear Enterprises (G.B.) Ltd, Sighthill, Edinburgh, Schottland

(Abbildung Seite 551)

Der Spectrascope-Kanal-Analysator Modell 100 ist das neuste, nunmehr aus dem Laben-Programm für kerntechnische Spektrometrieanalysensysteme der Nuclear Enterprises (G.B.) Ltd lieferbare Gerät.

Das tragbare, äusserst genaue und anpassungsfähige Gerät besteht aus Verstärker, Umsetzer, Speicherlogik, Decodiereinheit für Druckwerk und Mehrfachuntersetzer, arbeitet sowohl in Addition wie Subtraktion und hat Adressen- und Inhaltsregister, die in Serie—oder auf Wunsch in parallel—zugänglich sind. Das Instrument kann in folgenden Betriebsarten arbeiten: Statistische Erfassung, Mehrfachuntersetzung, Sampling, Mittelwertbestimmung und Impulshöhenanalyse. Der Analysator kann als unabhängiges Einkanalgerät mit Mehrfachuntersetzung betrieben werden und hat einen eingebauten unüberlastbaren Verstärker mit RC-Begrenzung. Man kann ihn mit einem Seriendruckwerk, auf Wunsch aber auch mit einem Paralleldruckwerk betreiben. Technische Kurzdaten sind: Zählkapazität 10^5 —1, Digitierungsrate 10 MHz, Schaltzeit 1,5 μs, Speicherzyklus 6 μs und eine Gesamttotzeit von 17 μs.

Der Einschub Modell 100A ist entweder als Zusatz zum Grundgerät oder unabhängig einzusetzen. Dieses Gerät enthält eine Hochspannungsversorgung für kerntechnische Tastköpfe, einen Programmzeitschalter für die Vorwahl der Mess- und Wartezeiten sowie einen Kanalwähler für Datenintegration und Selektion von Ausgangssignalen für automatische Stabilisation gegen externe Geräte.

EE 84 754 *für weitere Einzelheiten*

Digitalvoltmeter

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 551)

Dawe Instruments Ltd hat ein preisgünstiges Digitalvoltmeter und einen Wechselspannungs - Gleichspannungs - Konverter mit Typenbezeichnung 652 bzw. 653 angekündigt. Man kann sie entweder getrennt oder zusammen einsetzen, um bei 4stelliger Anzeige Wechsel- oder Gleichspannungen mit 0,1 oder 0,2 Prozent Genauigkeit zu messen.

Das Digitalvoltmeter 652 ist in den zwei Grundausführungen 652 A und 652 AX erhältlich. In der ersteren liefert eine einfache Normalzelle die Bezugsspannung für Messungen bis zu 0,2 mV herunter mit einer Gesamtunsicherheit von ± 0,2 Prozent des Bereichendwertes. Modell 652 AX hat eine Zenerdioden-Bezugsquelle und misst bis zu 0,1 mV herunter mit einer Gesamtunsicherheit von ± 0,1 Prozent des Endwertes. Beide Ausführungen haben vier Bereiche mit Endwertanzeigen von 1, 10, 100 oder 1 000 V— auf mechanischen Zählern mit Kommaautomatik. Die Eingangsimpedanz von 2,2 MΩ ist hoch genug, um Belastung der meisten Schaltungen zu vermeiden.

Die Konstruktion beruht auf einem selbstabgleichenden Servosystem; die Eingangsgleichspannung wird zusammen mit einer mittels motorgetriebenem Potentiometer geregelten Bezugsgleichspannung in einen Vergleichsverstärker gespeist. Der Vergleichsverstärker gibt eine Wechselspannung ab, die in Amplitude und Phase der Differenz zwischen Signal- und Bezugspegeln entspricht und in die Steuerwicklung eines Zweiphasen-Elektromotors gespeist wird. Der Motor treibt die Anzeigezähler und verschiebt den Potentiometerschleifer, bis die Bezugspannung der Eingangsgleichspannung des Vergleichers entspricht.

Die durchschnittliche Einstellzeit der Anzeige ist 2 s und höchstens 5 s für Höchstwerte. Der Nullpunkt kann unterdrückt werden. Das Instrument kann an das 50- oder 60-Hz-Netz angeschlossen werden und ist in ein kompaktes, tragbares blaues Metallgehäuse eingebaut, das robust genug für Werkstatteinsatz ist. Die Abmessungen sind 190 × 152 × 280 mm hoch, das Gewicht etwa 4 kg.

Der Wechsel-Gleichspannungskonverter 653 ermöglicht genaues Messen von Wechselspannungen mit dem Digitalvoltmeter 652. Seine Schaltung besteht aus einem Eingangsabschwächer, Verstärker mit hohem Verstärkungsgrad und Endstufe mit Filter für die Mittelwertbildung. Der Konverter kann mit Eingangssignalen vom Millivoltpegel bis zu 500 V_{eff} bei Frequenzen von 50 Hz... 25 kHz beaufschlagt werden. Der Eingang kann erdfrei sein und gibt maximal 1 V— für die drei niedrigsten Bereiche und 0,5 V— für den höchsten Bereich. Die Nenngenauigkeit zwischen 50 Hz und 10 kHz ist ± 0,3 Prozent des jeweiligen Bereichwertes und bis zu 25 kHz innerhalb ± 0,4 Prozent; die Linearität ist in allen Bereichen 0,05 Prozent. Das Gerät hat Netzanschluss und ist in einem dem Modell 652 angepassten Gehäuse untergebracht.

EE 84 755 *für weitere Einzelheiten*

Fernsprechkabel-Prüfgerät

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Abbildung Seite 551)

Ein tragbares, in sich geschlossenes Fernsprechkabel-Prüfgerät, das Installations-Ingenieuren wesentliche Vorteile bietet, wurde von Standard Telephones & Cables Ltd angekündigt.

Das mit Typ 74226 bezeichnete Gerät eignet sich für das Messen der Gegenkapazität, Kapazitätsunsymmetrie und prozentualen Widerstandsunsymmetrie. Man kann daher die Kapazitätsunsymmetrie in den verschiedenen Kombinationen messen, die in einzelnen Kabelabschnitten mit Vierer- und Paarverseilung vorkommen, so dass die Kabelabschnitte zur Kleinsthaltung von Nebensprechen und Störung in optimaler Anordnung gespleisst werden können.

Es gibt folgende Messbereiche:
Kapazitätsunsymmetrie

Bereich 1	280-0-280 pF
Bereich 2	1100-0-1100 pF

mit ±(1 Prozent +2 pF) Messunsicherheit. Gegenkapazitäten bis zu 0,1121 μF können mit ±50 pF Unsicherheit gemessen werden, und bei einer Widerstandsunsymmetrie von ±0,5 Prozent des Aderschleifenwiderstandes ist die Messgenauigkeit 0,01 Prozent.

Das Gerät kann zwecks besserer Beobachtung schräggestellt werden, und für Verwendung unter schlechten Lichtverhältnissen gibt es eine Skalenbeleuchtung gegen Preiszuschlag. Die bei Unsymmetrie der Messbrücke in den Kopfhörer gespeiste Tonfrequenzleistung ist begrenzt, um das Bedienungspersonal nicht zu irritieren.

Die mit einem transistorisierten Oszillator und Verstärker betriebene Brückenschaltung ist in zwei Ausführungen erhältlich, und zwar für 800 bzw. 1000 Hz. Als Stromversorgung dienen zwei eingebaute 4,5-V-Trockenbatterien.

Das Gerät wird in einem Holzgehäuse mit Traggriffen und abnehmbarem Deckel geliefert. Die Aussenabmessungen sind 432 × 305 × 254 mm, das Gewicht 19,5 kg.

EE 84 756 *für weitere Einzelheiten*

Oszillograf

Roband Electronics Ltd, Charlwood Works, Lowfield Heath Road, Charlwood, Horley, Surrey

(Abbildung Seite 552)

Der neue Oszillograf R0501 ist mit einer 127-mm-Oszillografenröhre bestückt, deren Betriebsspannung von 3,3 kV stabilisiert ist, um bei allen Schreibgeschwindigkeiten grosse Helligkeit zu erzielen. Es wird nur eine Druckschaltungsplatte benutzt, auf der alle Hauptbauelemente kodiert sind.

Das Gerät hat eine Bandbreite von 0...6 MHz, eine Anstiegszeit von 55 ns, Ablenkkoeffizienten von 50 mV/cm...20 V/cm in neun geeichten Bereichen (1-2-5-Stufen) und eine Messunsicherheit von ±5 Prozent.

Die Zeitablenkung ist 1 µs/cm bis zu 100 ms/cm in sechs geeichten Stufen; ein kontinuierlich im Verhältnis 10:1 regelbares Bedienelement erweitert den Bereich auf 1 s/cm; ein regelbarer x5-Vergrösserer erlaubt Höchstzeitablenkungen von 200 ns/cm. Eine vielseitige Triggerschaltung gestattet Wahl des Triggerpunktes, und ein eingebauter Fernsehhalbbild-Integrator ermöglicht Triggern der Zeitablenkung durch Fernsehhalbbildimpulse. Empfindlichkeit: intern 5 mm, extern 4 V_{SS}.

Ein interner Eicher ist vorhanden, und der Zweifarbenraster hat Flutbeleuchtung; das Gerät kann mit Standardkameras für 5" (127 mm) Mittellinienabstand benutzt werden. Die Gesamtabmessungen sind 251 mm hoch, 375 mm tief und 191 mm breit, das Gewicht ist etwa 8 kg.

EE 84 757 *für weitere Einzelheiten*

Drahtschneider und -abisolierer

Multicore Solders Ltd, Multicore Works, Maylands Avenue, Hemel Hempstead, Hertfordshire

(*Abbildung Seite 552*)

Der neue Drahtschneider und -abisolierer "Bib" Modell Acht ist mit einer vorwählbaren Einstellung auf Drahtstärken von 0,46 bis 2,65 mm ausgerüstet. Bei richtiger Einstellung des Vorwählers wird die Isolierung ohne Beschädigung des Drahtes abgestreift. Das Werkzeug hat gepolsterte Plastikhandgriffe, um das Greifen zu erleichtern.

EE 84 758 *für weitere Einzelheiten*

Digitalvoltmeter

Radun Controls Ltd, White Street, Caerphilly, Glamorganshire

(*Abbildung Seite 552*)

Radun Controls Ltd kündigt zwei neue Digitalvoltmeter DM 850 und DM 851 für universelle Verwendung in technischen Labors von Industrie und Lehranstalten an.

In der Konstruktion beider Messgeräte, die eine vierziffrige Anzeige mit 0,1 Prozent Genauigkeit für Gleichspannungen und 0,3 Prozent für Wechselspannungen geben, wurden ohne Einbusse an Leistung oder Betriebssicherheit wesentliche Kosteneinsparungen erzielt.

Das abgebildete Universalvoltmeter DM 850 hat einen Messumfang von 0...1090 V≃ in vier Bereichen mit Endwerten 1,090 V, 10,90 V, 109,0 V bzw. 1090 V und von 0...109 mA mit 109,0 µA, 1,090 mA, 10,90 mA bzw. 109,0 mA. Das Auflösungsvermögen beider Geräte ist 0,02 Prozent des Bereichendwertes, die Eingangsimpedanz 100 MΩ im Bereich 1 V— und 30 MΩ mit 20 pF Nebenschluss im Bereich 1 V~; der Eingangswiderstand der anderen Bereiche ist 2 MΩ bei einem Betriebsfrequenzbereich des Gerätes von 20 Hz bis zu 50 kHz. Eingangssignale können einer erdfreien oder geerdeten Quelle entnommen werden, und beide Voltmetermodelle sind gegen unbeabsichtigte Überlastung geschützt.

Beide Voltmeter sind mit Ausnahme des selbstabgleichenden Servos durchweg in Festkörperschaltungen in Form von Chopper-Verstärkern mit Driftkompensation ausgeführt und mit Präzisionsbezugsspannung und Wechsel-Gleichspannungskonverter ausgerüstet.

EE 84 759 *für weitere Einzelheiten*

Elektronische Zeitschalter

Controls & Automation Ltd, Regal House, 55 Bancroft, Hitchin, Hertfordshire

(*Abbildung Seite 552*)

Diese Geräte wurden speziell als Alternative zu den Zeitschaltertypen mit Synchronmotor und Trigger entwickelt; sie arbeiten mit einer Fadentransistor-Verzögerungsschaltung, die ein Starkstromrelais ansteuert. Sie können entweder in Chassisausführung für Einbau oder Direktmontage an die Maschine oder in einem kompakten Stahlgehäuse geliefert werden.

Lieferbare, durch integrale oder ferngesteuerte Potentiometer einstellbare Zeitbereiche sind von 0,05...1 s bis zu 0,25...60 s. Die Schaltgenauigkeit von ±1 Prozent wird durch Netzschwankungen von ±10 Prozent nicht beeinflusst. Die Instrumente können so angeordnet werden, dass sie während oder am Ende der Zeitperiode die externe Schaltung betätigen.

EE 84 760 *für weitere Einzelheiten*

HF-Leistungsmonitor

Vertrieb: Wessex Electronics Ltd, Royal London Buildings, Baldwin Street, Bristol, 1

(*Abbildung Seite 552*)

Der neue, zweiseitig gerichtete Leistungsmonitor Sierra 164 B zeigt die einfallende oder reflektierte Leistung von 50 koaxialen Systemen kontinuierlich oder aussetzend während der Übertragung an. Antennen oder Blindlasten lassen sich Sendern oder anderen Leistungsquellen einfach durch Überwachen der während des Betriebes reflektierten Leistung mit dem Modell 164 B genau anpassen. Das Umschalten von reflektierter auf einfallende Leistung erfolgt durch Drehen eines einzigen Knopfes oben am Gerät. Mit jedem Einsatz gibt es vier Bereiche. Zehn Einsteck-Elemente gestatten Wahl von Frequenz- und Leistungsbereichen von 2...1 000 MHz und von 100 mV...5 kW. Die Leistung wird direkt von einer linearen Skala abgelesen, und es gibt auch eine Skala für Stehwellenverhältnisse. Hohe Richtfähigkeit und ein niedriges Einfügungsstehwellenverhältnis sichern maximale Genauigkeit bei geringster Störung der zu prüfenden Übertragungsleitung. Das Sierra-Messwerk ist gedämpft, so dass die Lage beim Einsatz gar keine Rolle spielt, was den Transport des Gerätes erleichtert. Modell 164 B ist mit den neuen "Abdreh-Steckverbindungen" der Sierra oder Kundenwünschen entsprechend lieferbar.

EE 84 761 *für weitere Einzelheiten*

Digitalvoltmeter

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(*Abbildung Seite 553*)

Ein komplettes neues Programm für preiswerte Digitalthermometer, die ohne Schritthalter oder Relais in beiden Richtungen nachlaufen können, wurde von Wayne Kerr Laboratories Ltd angekündigt. Eine beleuchtete 4stellige Anzeige, deren letzte Ziffer kontinuierliche Änderungen darstellt, gibt ein hohes Auflösungsvermögen. Vollaustauschbare Thermistor-Messfühler sind in 17 Ausführungen für Labors, medizinische, meteorologische, industrielle, biologische und chemische Verwendungszwecke lieferbar. Mann kann bis zu vier Messfühler zur gleichen Zeit an das Gerät anschliessen und sie einzeln über ein Bedienelement auf der Frontplatte einstellen.

Alle Thermometer zeigen direkt in Fahrenheit oder Celsius an und sind entweder mit grossem Messumfang oder —für besondere Anwendungszwecke— mit 5stelliger Anzeige für einen bestimmten Messumfang von 20° oder 40° F zu haben. Die Festkörperschaltung kann ohne Schwierigkeiten für Abgabe eines binär-kodierten Dezimalsignals modifiziert werden; ausserdem gibt es noch wahlweise ein Weitergabepotentiometer für externes Untersetzen.

Das Thermometer ist in dem gleichen kompakten Gehäuse untergebracht wie das Digitalvoltmeter (203 × 152 × 152 mm). Einzelheiten über Platin-Ausführung werden in Kürze bekanntgegeben.

EE 84 762 *für weitere Einzelheiten*

Überschall-Reinigungsgerät

Kerry's (Ultrasonics) Ltd, Warton Road, London, E.15

(*Abbildung Seite 553*)

Kerry's (Ultrasonics) Ltd gibt bekannt, dass sie in Zusammenarbeit mit Deltasonics Inc., Hawthorne, Kalifornien, USA, ihre "Pulsatron"-Überschallreinigungsgeräte herausbringen. Der Überschallgenerator KG.850 hat nur ein Bedienelement mit Positionen für hohe, mittlere und niedrige Leistung und ein Volumen von nur etwa 1 500 cm^3. Der Generator ist volltransistorisiert und hat eine eingebaute patentierte Einrichtung, die unabhängig von den Wandlern automatisch mitläuft und Einsatz von Wandlern in Sandwichkonstruktion mit hohem

Q und niedriger Impedanz gestattet. Durch diese Anordnung werden "tote Punkte" vermieden und die gesamte Energie in der Wanne zerstreut. Der Generator erfordert keine Anheizzeit, stellt sich automatisch auf das Flüssigkeitsniveau in der Wanne, die zu reinigende Last oder Eingangsspannung ein und wird weder durch Leerlauf- noch Kurzschlussbedingungen beeinflusst. Da alle Schaltungen in Festkörpertechnik ausgeführt sind, wird höchste Zuverlässigkeit und störungsfreier Betrieb gewährleistet. Der Generator wiegt ca. 16 kg.

Die Wanne für die Überschallbehandlung ist gänzlich aus poliertem rostfreien Stahl konstruiert und mit einer Abflussvorrichtung versehen. Die Innenabmessungen und aktive Fläche der Wanne sind 355 × 305 × 305 mm. Die an den Boden der Behandlungswanne angebauten modularen "Pulsatron"-Wandlerelemente arbeiten nach dem elektrostriktiven Prinzip und sind in Sandwichkonstruktion aus Aluminiumblöcken und Bleizirkonattitanat-Keramik aufgebaut. Die auf den Boden montierten modularen "Pulsatron"-Wandler können Flüssigkeitstemperaturen bis zu etwa 150°C aushalten, was für alle herkömmlichen Reinigungen mit Ultraschall genügt.

Diese Baureihe von Überschall-Reinigungsgeräten ist für Betrieb bei einer Frequenz von 25 kHz oder 45 kHz konstruiert, jedoch werden die Geräte der ersten Fertigungsserie für Betrieb bei 25 kHz angeboten. Diese Entwicklung geht der Einführung eines Überschall-Reinigungssystems mit 2 kW mittlerer und 4 kW Spitzenleistung voraus. Patente sind angemeldet.

EE 84 763 *für weitere Einzelheiten*

Empfänger-Schutzgerät

Racal Communications Ltd, Western Road, Bracknell, Berkshire

(*Abbildung Seite 553*)

Racal Communications Ltd hat vor kurzem ihr Programm für HF-Verkehrsaustüstungen durch den Typ MA.197 für HF-Empfängervorselektion und -schutz ergänzt.

Der MA.197 wurde für erfolgreichen Empfängerbetrieb in unmittelbarer Nähe von Hochleistungssendern und ähnliche ungünstige Bedingungen, die Herabsetzung des Durchbruchs sowie des mechanischen und elektrischen Empfängerschadens erfordern, entwickelt. Diese Situation ist für Schiffsanlagen unvermeidlich und besteht auch dort, wo Sender und Empfänger im selben Gebäude oder in transportablen Behältern installiert sind.

Durch Einschalten des MA.197 zwischen Antenne und Eingangsschaltungen des Empfängers lässt sich eine scharfe Vorselektion erreichen, wodurch Nachbarkanalstörungen und Kreuzmodulation reduziert werden. Für ein ungewünschtes, um 5 Prozent verstimmtes Signal wird eine Abschwächung von über 95 dB erreicht, und zwar ohne Herabsetzung der Gesamtverstärkung und bei nur sehr kleiner Empfindlichkeitseinbusse. Ausserdem schützt der MA.197 den Empfänger gegen Ausbrennen durch übermässige HF-Eingangssignale, da er ohne Schaden Beaufschlagung mit EMKs bis zu 40 V aushält.

Bei Vorselektion und vollem Überlastungsschutz wird der MA.197 entweder an Ort und Stelle oder durch Fernbedienung mittels Einzelknopf über den Bereich 2...24 MHz abgestimmt; wo nur Überlastungsschutz erforderlich ist, kann von 1...30 MHz abgestimmt werden. Das Gerät hat Netzanschluss für alle herkömmlichen Wechselstromnetze und nimmt ca. 25 W auf. Es hat eine Standard-19"-Frontplatte für Gestelleinbau und ist für Quer-Druckluftkühlung geeignet.

EE 84 764 *für weitere Einzelheiten*

Urraster für gedruckte Schaltungen

Krisson Printing Ltd, Equipment Division, 184 Acton Lane, London, N.W.10

(*Abbildung Seite 553*)

Der Geschäftsbereich Ausrüstungen der Krisson Printing Ltd bietet eine Auswahl beleuchteter Tische mit eingebautem Krisson-Urraster für die Erstellung von Druckschaltungsoriginalen an.

Für gedruckte Schaltungen wird der Urraster mit 0,1" (2,54 mm) Grundmass geliefert, wobei die lineare Toleranz über 1016 mm ±0,1 mm (über 508 mm ±0,05 mm) ist. Diese permanenten, 6,35 mm dicken Glasraster gestatten schnelles und genaues Zeichnen. Die Beleuchtung von unten zeichnet die Rasterlinien scharf und klar für momentane Führung auf der Schaltung ab.

Für sofortige Kennzeichnung und Registereinstellung für die gedruckten Schaltungen kann der Urraster mit einem Stift- und Fahnenregistriersystem geliefert werden. Der Urraster ist nahe jeder der beiden kurzen Seiten der Glasoberfläche mit einem kleinen Stift ausgestattet; transparente Fahnen kommen in Paketen von je 50 Stück.

Jedes transparente Original wird lose über die Stifte gelegt. Über jeden Stift wird dann eine genau gelochte Registerfahne gelegt und mit Klebstreifen auf die Transparenz geklebt. Dieser Vorgang wird mit jeder weiteren Pauspapierlage wiederholt. Der gesamte Satz ist dann mit den Glasrasterlinien ausgerichtet, und jedes Blatt ist in genauem Register mit jedem anderen.

EE 84 765 *für weitere Einzelheiten*

Klirrfaktormesser

Dymar Electronics Ltd, Rembrandt House, Whippendall Road, Watford, Hertfordshire

(*Abbildung Seite 554*)

Das Dymar-Einschubsystem für Messgeräte wurde durch den Klirrfaktormesser 765 ergänzt. Der Einschub gestattet Messen des Gesamtklirrfaktors einer Tonfrequenz im Bereich 20 Hz ... 20 kHz.

Das zugrunde liegende Prinzip ist die Unterdrückung der Grundkomponente der angelegten Welle und das Messen des Rückstandes. Die Voltmeterbereiche sind direkt in Klirrfaktor geeicht, und der Messumfang erstreckt sich von 100 Prozent bis zu 0,025 Prozent herunter. Die Messergebnisse können auch in Dezibel unter der Grundkomponente ausgedrückt werden. Die Tieffrequenzen kann man ausfiltern, um Netzfrequenzkomponente zu entfernen.

EE 84 766 *für weitere Einzelheiten*

Miniaturrelais

Magnetic Devices Ltd, Exning Road, Newmarket, Suffolk

(*Abbildung Seite 554*)

Magnetic Devices Ltd hat neue preiswerte Miniaturrelais eingeführt, bei deren Konstruktion vor allem auf die Herabsetzung der Anzahl zusammenzubauender Teile Wert gelegt wurde. Das Relais hat eine transparente Haube aus Makrolon, ein Werkstoff, der nicht brennt, und ist direkt auf den Sockel montiert, so dass keine getrennten Befestigungswinkel benötigt werden. Ausserdem wird der Relaiskörper zur weiteren Stützung in Nuten der Haube geführt.

Durch diese neuen Relaiseigenschaften wird sowohl die Höhe als auch die Einbaufläche wesentlich verringert und durch Reduktion der Anzahl der Verbindungspunkte die Zuverlässigkeit erhöht. Besondere Aufmerksamkeit wurde Kriechströmen und Lufträumen zwischen elektrisch isolierten Teilen gewidmet, wodurch selbst bei schwierigem Schalten lange Lebensdauer erzielt wird.

Baumuster 120 für Wechselstrom und 125 für Gleichstrom sind zweipolige Umschaltrelais, für 6 A bemessen und fest auf einen gepressten 8poligen Stecksockel genietet. Baumuster 120 hat eine Nennleistung von 3 VA, eine höchste Betriebsspannung von 250 V und einen maximalen Standardspulenwiderstand von 7 300 Ω.

Baumuster 130 (Wechselstrom) und 135 (Gleichstrom) sind üblicherweise mit 3poligen Umschaltkontakten bestückt und auf einen gepressten 11poligen Stecksockel mit ausreichenden Stegen zur Verhütung von Kriechwegbildung montiert. Bei 6 A Nennstrom hat das Baumuster 130 eine Nennleistung von 3,5 VA, Höchstbetriebsspannung von 250 V und einem Höchstspulenwiderstand von 7 300 Ω. Baumuster 135 hat eine Nennleistung von 1,5 W, Höchstbetriebsspannung von 125 V und Höchstspulenwiderstand von 7 300 Ω.

Die Arbeitszeit der Baumuster 125 und 135 ist 10 ms, die Abfallzeit 6 ms. Die neuen Relais arbeiten in Umgebungstemperaturen bis zu 40°C, können im Sonderauftrag jedoch für höhere Temperaturen modifiziert werden.

Das Höchstgewicht ist 85 g. Die Relais können mit Feinsilber- oder Elkonitkontakten, sowie für feuchte oder tropische Bedingungen vakuumimprägniert geliefert werden.

EE 84 767 *für weitere Einzelheiten*

Optischer Drehzahlgeber

Ferranti Ltd, Thornybank, Dalkeith, Scotland

Der optische Drehzahlgeber von Ferranti erzeugt elektrische Impulse, deren Frequenz der Wellengeschwindigkeit des treibenden Mechanismus proportional ist. Die Impulshöhe ist unabhängig von der Drehzahl.

Das Gerät besteht aus einem Muster radialer Linien auf einer Glasscheibe, die auf die Eingangswelle montiert ist. Licht von einer Glühfadenlampe wird kollimiert und fällt durch die Scheibe auf eine Gruppe von Fotoelementen. Durch Verwendung einer Gruppe von Fotoelementen wird der Einfluss von Schwankungen der Lichtquelle verringert. In den Lichtstrahl eingeschoben ist ein kleines Glasstück mit einem Linienmuster, das dem auf der Scheibe ähnlich ist. Einstellen der Linien auf dem kleinen Glasstück in solcher Weise, dass sie mit den Linien der Eingangsscheibe einen Winkel bilden, ruft Mohrsche Streifenbilder hervor. Die Streifenbilder bewegen sich über die Fotoelemente, wenn die Scheibe rotiert und erzeugen in diesen die erforderlichen Ausgangsimpulse.

Zur Drehrichtungsermittlung oder Frequenzverdopplung können zwei, drei oder vier Fotoelemente vorgesehen werden, und die Phasenbeziehung der Ausgangssignale der Sperrschicht-Fotozellen zueinander ist etwa 0°, 90°, 180° und 270°. Interne Verstärker zur Verstärkung dieser Signale sind lieferbar.

Eine 3 V, 3 W Glühfadenlampe dient als Lichtquelle und wird nur mit 2,4... 2,5 V betrieben, so dass die Lebensdauer mindestens 1000 Stunden beträgt. Die Lampe lässt sich ohne Ausbau des Verschlüsslers auswechseln.

In der vorliegenden Konstruktion ist das Instrument für insgesamt 100 Impulse/Umdrehung/Fotoelement ausgelegt, was jedoch mittels elektronischer Schaltungen auf 200, 400 oder 1000 erhöht werden kann.

Weitere Modelle sind in der Entwicklung, und die zur Zeit in Betracht gezogenen Impulszahlen/Umdrehung sind 100, 180, 250, 256, 375, 540, 720, 900, 1000, 1080, 1600, 2048, 2250 und 2700.

Ausserdem könnte man in Zukunft das Programm durch Modelle für niedrigere Anzahlen ergänzen, z.B. für 10, 20, 25, 30, 40, 50 und 60 Impulse/ Umdrehung.

Die zwei Hauptanwendungsmöglichkeiten für die optischen Drehzahlgeber sind einmal als Tachometer, da die Ausgangsfrequenz als Drehzahl angegeben werden kann, oder zum Messen von Wellenwinkeln, die man durch Zählen der Impulse von einem vorgewählten Bezugspunkt bestimmen kann.

In der letzteren Anwendungsmöglichkeit ist das Gerät für Zählen in beiden Richtungen geeignet, z.B. für das Positionieren von Werkzeugmaschinen entlang der X- und Y-Achse, wenn der Geber von der Leitspindel oder einer beliebigen Welle, deren Winkelbewegung der zu steuernden oder zu messenden Bewegung proportional ist, getrieben wird.

EE 84 768 *für weitere Einzelheiten*

UKW-Sende-Empfangsgerät

Radio Communications Co., 16 Abbey Street, Crewkerne, Somerset

(Abbildung Seite 554)

Der UKW-Funkfernsprecher "Telecomm" TRT/4 ist ein transportables und volltransistorisiertes Gerät für Gegensprechverkehr zwischen ähnlichen Geräten oder ortsfesten und beweglichen Stationen. Es ist für Ein- oder Zweifrequenzbetrieb auf ein bis sechs Kanälen im Frequenzbereich 50...100 MHz ausgelegt.

Durch vollständige Bestückung des Sender- und Empfängerteils mit Transistoren wurde die Leistungsaufnahme auf einen im Vergleich mit Röhrenausrüstungen ungewöhnlich niedrigen Wert reduziert. Eine interne dichte und aufladbare Batterie gestattet bei einem Empfangs-Sendeverhältnis von 10 : 1 ungefähr neun Betriebsstunden. Die Batterie lässt sich schnell ohne Öffnen des Gerätes und ohne Werkzeuge ausbauen. Auf Wunsch kann sie auch im Gerät über das externe Ladegerät BC/4 aufgeladen werden. Das Gerät kann auch aus einer externen Stromversorgung für 28 V—, z.B. einer Hochleistungsbatterie oder einem Wechselstrom-Netzgerät gespeist werden.

Eine einstellbare Peitschenantenne (1/4 Wellenlänge) wird benutzt, und eine koaxiale Steckdose gestattet Anschluss externer Antennensysteme oder Testgeräte. Ein Tauchspulenlautsprecher ist eingebaut.

Ein Handmikrofon in Kunststoffgehäuse mit Sprechtaste gestattet einfachen und schnellen Verkehr. Bei Empfang kann dieses Mikrofon auch als kleiner Lautsprecher dienen. Auf Wunsch ist auch statt des Handmikrofons ein leichter Sprechhörer erhältlich. Ausserdem gibt es auch eine Doppelkopfhörergarnitur mit Bügelmikrofon.

EE 84 769 *für weitere Einzelheiten*

Universal-Messbrücke

Thomas Industrial Automation Ltd, Deansgate Lane, Altrincham, Cheshire

(Abbildung Seite 555)

Dieses Instrument ist für das Messen von Induktivität, Kapazität und Widerstand mit ±1 Prozent Unsicherheit ausgelegt. Es ist völlig in sich geschlossen, transistorisiert und wird aus eingebauten Batterien betrieben.

Die Brücke kann mittels ihres internen Oszillators bei 100 Hz, 1 kHz und 10 kHz erregt werden. Zum Messen des Widerstandes von Bauelementen mit Scheinwiderstand, z.B. Transformatoren mit Eisenkern, kann die Brücke durch Gleichstrom—wiederum aus internen Batterien—erregt werden.

Als Nulldetektor dient ein Drehspulmesswerk, dem ein Verstärker mit hoher Verstärkung vorgeschaltet ist. Zur Erleichterung des Abgleichens der Brücke mit Induktoren und Kondensatoren mit niedrigem Leistungsfaktor ist eine automatische Verstärkungsregelung vorhanden. Bei Gleichstrom wird ein Transistor-Chopper eingeschaltet, der den Verstärker in einen empfindlichen Gleichstromdetektor konvertiert.

EE 84 770 *für weitere Einzelheiten*

X-Band-Miniaturzirkulator

Microwave Systems Ltd, 9-10 River Front, Enfield, Middlesex

(Abbildung Seite 555)

Microwave Systems Ltd hat ihr umfangreiches Sortiment von Micromega-Bausteinen in Grossbritannien durch einen neuen koaxialen Miniaturzirkulator-Isolator für das X-Band ergänzt.

Der im Bereich 8,5 ... 9,5 GHz mit 25 dB Trennung arbeitende Zirkulator-Isolator hat einen Einfügungsverlust von 0,4 dB und ein Stehwellenverhältnis von 1,15. Die Abmessungen des Bausteins sind 25,4 mm × 28,6 mm × 19 mm ausschliesslich Anschluss, das Gewicht etwa 115 g einschliesslich N-Anschluss.

Ähnliche Zirkulatoren-Isolatoren sind auch für die L-, S- und C-Bänder lieferbar.

EE 84 771 *für weitere Einzelheiten*

Transistor-Kennlinienschreiber

Radar & Communication Instruments Division, Elliott-Automation Ltd, 2-4 Wigton Gardens, Stanmore, Middlesex

(Abbildung Seite 555)

Mit dem vom Geschäftsbereich Radar- und Nachrichteninstrumente der Elliott-Automation hergestellten neuen Transistor-Kennlinienschreiber Elliott 8100 lassen sich die Kennlinien eines breiten Sortimentes von Halbleitern darstellen.

Das kompakte Instrument, das in gleicher Weise für Entwicklungsarbeiten im Labor, Testen in der Fertigung oder als Lehrmittel in Unterricht und Ausbildung geeignet ist, enthält einen speziell entwickelten Oszillografen sowie Wellenformgeneratoren, die für das Testen von Halbleitern unter besonderen Bedingungen erforderlich sind.

Man kann sowohl npn- wie pnp-Transistoren messen und ähnliche Halbleiter schnell auf Zusammenpassen vergleichen.

Das neue Gerät hat zwei Grunddarstellungsweisen: Ansteuerung des zu prüfenden Halbleiters mit einer Treppenwellenform führt zur Darstellung einer Kurvenschar, oder ein Einstufeneingang mit kontinuierlich regelbarer Amplitude kann den gesamten Strom- und Spannungsbereich überstreichen, was besonders beim Testen von Thyristoren nützlich ist.

Folgende Kennlinien sind darstellbar:

I_c als Funktion von V_{ce} mit I_b oder V_{be} als Parameter

I_c als Funktion von V_{be}

I_c als Funktion von I_b

I_{ceo} als Funktion von I_{ces};

ausserdem Durchlass- und Sperrkennlinien von Zener-Dioden, Tunneldioden und Gleichrichtern.

Diese Darstellungen sind sehr informativ, vor allem in bezug auf die Lawinen- und Durchgriffspannungen von Transistoren und Durchbruchkennlinien von Vierschichtdioden und Thyristoren.

Genaue Strom- und Spannungsquellen sind eingebaut, so dass der Benutzer die Darstellung vor Durchführung der Messungen eichen kann.

EE 84 772 *für weitere Einzelheiten*

Hochbelastbare Stromversorgung

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Abbildung Seite 555)

Gresham Lion Electronics Ltd hat eine hochbelastbare kombinierte elektronische Stromversorgung und Ladeausrüstung herausgebracht. Das luftgekühlte Gerät gibt bei Anschluss an 200 V, 400 Hz Drehstrom bei konstanter Spannung von 28 V (±1 V) einen Höchststrom von 90 A— ab. Die Ausrüstung hat die Typenbezeichnung GLE 400 und wurde vom Britischen Verteidigungsministerium für die Streitkräfte gewählt.

Das Modell GLE 400 gibt für das Laden von Batterien Konstantstrom in Stufen von 15 A bis zu 75 A (±10 Prozent) bei Spannungen zwischen 10 und 31 V ab. Die Restwelligkeit ist unter allen Bedingungen niedriger als 0,5 V_{eff}.

Die Ausrüstung kann beide Funktionen zur gleichen Zeit erfüllen, vorausgesetzt dass die Summe der beiden Ausgängen entnomenen Ströme 90 A nicht überschreitet.

Die Aussenabmessungen des für ortsfesten oder fahrbaren Einsatz bei Temperaturen zwischen −20°C und +50°C und Feuchtigkeiten von 0...90 % geeigneten Gerätes sind ca. 40 cm × 40 cm × 91 cm, das Gewicht etwa 96 kg.

Schmelzsicherungen im Ein- und Ausgang sowie ein thermostatischer Sicherungsautomat verhindern Schaden bei Überlastung oder Überhitzen durch Einschränkung des Luftstroms oder Ausfallen des Luftgebläses.

Die Stromversorgung ist in einem robusten Aluminiumgehäuse mit vertieften Bedienfeldern und Traggriffen untergebracht. Die Lufteinlassöffnungen sind besonders gestaltet, um Schaden durch Wasserspritzer zu verhindern.

Alle elektronischen Regelschaltungen sind auf drei Einsteck-Druckschaltungen angeordnet, wodurch Wartung und Service wesentlich vereinfacht werden.

EE 84 773 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein Verstärker mit konstanter Lautstärke über einen grossen dynamischen Bereich von A. G. Morris

Zusammenfassung des Beitrages auf Seite 502-507

Fortschritte in der Konstruktion von Verstärkern mit konstanter Lautstärke wurden durch Abwesenheit eines idealen Regelelementes begrenzt. Der Transistor in Randwertschaltung kommt dem Ideal sehr nahe, da er sehr niedrige Verzerrungen hervorruft und einen denkbaren Regelbereich von über 120 dB hat. Ein System mit Transistor in Randwertschaltung, das Eingangsschwankungen von 60 dB in 4 dB presst, wird beschrieben. Die Auswertung dieses breiten Dynamikumfanges bringt jedoch zwei weitere Probleme mit sich: erstens müssen die Bedingungen für den optimalen Konstanzspielraum eines Grossumfangregelsystems ausgearbeitet werden, und zweitens sind zusätzliche Schaltungen erforderlich, um die Verstärkung des Rauschens zwischen Worten zu blockieren. Beide Probleme werden gelöst; die Lösung des ersteren hat allgemeine Anwendungsmöglichkeiten in der Regeltheorie.

Mikrowellen-Tunneldiodenoszillator von J. Swift

Zusammenfassung des Beitrages auf Seite 508-511

Der Beitrag bespricht Anwendungsmöglichkeiten von Tunneldioden in Mikrowellenoszillatoren. Nach der Einführung werden die Parameter von Tunneldioden gegeben und daraufhin ihre Anwendung in Mikrowellen-Oszillatorschaltungen diskutiert. Abschliessend werden alle für den Entwurf eines 2-GHz-Oszillators erforderlichen Daten und ein Literaturnachweis für Veröffentlichungen auf diesem Gebiet gegeben.

Magnetische Elemente mit Vielfachöffnungen für Anwendung als rein magnetische Logik von D. J. Morris

Zusammenfassung des Beitrages auf Seite 512-519

Dieser Beitrag beschreibt die Eigenschaften einer Anzahl von magnetischen Elementen mit Vielfachöffnungen verschiedener Form, die für logische Operation in Grundverknüpfung entwickelt wurden. Zur Ausführung einander folgender logischer Operationen kann man die beschriebenen Elemente kombinieren, ohne dass zusätzliche Bauelemente ausser dem für das Verbinden verwendeten Draht erforderlich werden. Das für das Koppeln der Elemente angewandte System wird mit MAD-R bezeichnet. Es gestattet breite Betriebstoleranzen und ist für Anwendungszwecke gedacht, bei denen es auf höchste Zuverlässigkeit ankommt.

Die Leistung und Beschränkungen mit Tunneldioden bestückter Vorverstärker von B. Easter

Zusammenfassung des Beitrages auf Seite 520-523

Der Tunneldiodenverstärker hat Anwendungsmöglichkeiten als einfache Empfängereingangsstufe für Systeme über 1 GHz, für die ein mittelmässiger Rauschfaktor in Kauf genommen werden kann. Die Leistung von Verstärkern dieser Art wird in bezug auf Sichtlinien-Ballempfangssysteme untersucht. Nichtlineare Wirkungen, die auftreten können, einschliesslich AM-PM-Umwandlung, werden in Betracht gezogen.

Beim Entwurf eines praktischen Verstärkers auftretende Probleme—u.a. die Erzielung von Schaltungsstabilität—werden besprochen und Einzelheiten eines bei 2 GHz betriebenen Verstärkers gegeben.

Eine einfache Messschaltung mit Digitalanzeige für die Grossignalstromverstärkung von Transistoren von S. L. Hurst

Zusammenfassung des Beitrages auf Seite 524-528

Eine einfache Gleichstrom-Messschaltung wird beschrieben, die zusammen mit jedem herkömmlichen Frequenzmesser die Direktanzeige des Wertes der Grossignalstromverstärkung von Transistoren in Emitterschaltung ermöglicht. Die Messgenauigkeit entspricht der eines üblichen kommerziellen Gleichstrom-Transistortesters, hat aber den Vorteil leichter und eindeutiger Ablesung. Restströme können in Emitterschaltung gemessen werden, und das Aussuchen von Transistoren nach Stromverstärkung kann mit grosser Genauigkeit erfolgen.

Automatische Verstärkungsregelung für Rauschen von J. S. Bowman und H. Sutcliffe

Zusammenfassung des Beitrages auf Seite 529-531

Die Anwendung automatischer Verstärkungsregelung auf Rauschverstärker wird durch die ungeeigneten Betriebsdaten von Detektoren mit Verzögerung vor der Demodulation kompliziert. Eine Analyse des Frequenzganges solcher Detektoren wird für den Fall, dass das Rauschen normal verteilt ist, gegeben und daraus die Schlussfolgerung gezogen, dass für wirksame automatische Verstärkungsregelung Verstärkung und Verzögerung nach der Demodulation erforderlich sind. Eine Schaltung für die Regelung eines rauscherzeugenden Verstärkers im Bereich 10 . . . 100 kHz wird beschrieben. Das Regelelement ist ein Dioden-Spannungsteiler mit geregeltem Vorspannstrom.

Ein stickstoffgekühlter, entarteter parametrischer 6-GHz-Verstärker für eine radiometrische Anwendung von D. Chakraborty und G. F. D. Millward

Zusammenfassung des Beitrages auf Seite 532-535

Der Entwurf eines mit flüssigem Stickstoff gekühlten, entarteten parametrischen 6-GHz-Verstärkers für Anwendung in einem Radiometer für Antennengewinnmessungen einer Nachrichtensatelliten-Bodenstation über einen Radiostern wird beschrieben. Zwei Stufen von je 15 dB Verstärkung und 200 MHz Bandbreite bis zu den –3-dB-Punkten werden in Kaskade geschaltet und ergeben 30 dB Mindestverstärkung. Die dem Verstärker eigene Bandbreite reicht für diesen Zweck aus, und Sonderschaltungen zur Erweiterung der Bandbreite sind nicht erforderlich. Die Rauschtemperatur des gekühlten Verstärkers ist 57°K.

Die thermischen Empfindlichkeitsgrenzen eines Radiometersystems mit diesem Verstärker werden kurz besprochen.

Samplingsystem für hochgradige Konstanz von Gleichstrommotoren von F. Szlavik und G. Orbán

Zusammenfassung des Beitrages auf Seite 536-539

Ein System für vollautomatisches Anlaufen und hochgradige Geschwindigkeitskonstanz von Gleichstrommotoren niedriger und mittlerer Leistung wurde entwickelt. Die eingehend beschriebenen elektronischen Schaltungen wurden für einen Motor entworfen, der einen Neutronenstrahlzerhacker treibt. Die Genauigkeit und Konstanz von Frequenz und Phase werden für einen breiten Querschnitt beeinflussender Parameter mittels eines Bezugs-Messenders bestimmt.

Verzerrung und Ausgangsleistung von pulsbreitenmodulierten Verstärkern von E. C. Bell und T. Sergent

Zusammenfassung des Beitrages auf Seite 540-542

Die Pulsbreitenmodulationstechnik wird vom Gesichtspunkt ihrer Anwendung auf hochleistungsfähige lineare Tonfrequenzverstärker untersucht. Das Spektrum des modulierten Signals wird analysiert; aus den Ergebnissen wird gefolgert, dass Verzerrungen erwartet werden müssen und wie man dieselben durch Wahl einer geeigneten Abtastfrequenz auf ein Mindestmass reduzieren kann.

Es wird gezeigt, dass bei einer Tonfrequenzbandbreite von 15 kHz und Einflankenmodulation eine Abtastfrequenz von nicht weniger als 90 kHz erforderlich ist, wenn das Ausgangssignal keine Amplitudenkomponenten, die grösser als 1 Prozent des Tonfrequenzsignals sind, enthalten soll. Mit Doppelflankenmodulation kann die Abtastfrequenz auf 75 kHz reduziert werden.

Es wird weiterhin gezeigt, dass in der Praxis eine Ausgangsleistung in der Grössenordnung von 20 W das mit jederzeit greifbaren Transistoren wirtschaftlich erreichbare Maximum darstellt.

Eine reziproke Zeitintervalldarstellung mit Transistorschaltungen von R. H. Kay

Zusammenfassung des Beitrages auf Seite 543-545

Das beschriebene Gerät wurde für neurophysiologische Zwecke entwickelt und kann "momentane" Impulsfolgefrequenzen und ausserdem jeden Impuls, wenn er auftritt, registrieren. Der Ausgang wird auf dem Schirm einer Oszillografenröhre registriert und Impulsfolgefrequenzen von etwa 7 bis zu 700 sec^{-1} linear dargestellt.