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COMMENTARY

COLOUR television at last—or nearly so at least. The Postmaster General's statement made in the middle of February has received widespread publicity already and anything written now is, so to speak, for the record.

In effect, the Government has accepted and adopted the recommendations of the Television Advisory Committee which, in summary, are that an early start should be made on a change of standards, using the duplication method. This involves transmitting BBC 1 and independent television programmes on 625 lines on u.h.f. as well as maintaining, for a number of years, the present transmissions of these programmes on 405 lines on v.h.f. It was also recommended that colour should be introduced into all the 625 line transmissions.

In adopting these proposals the Government has claimed that, firstly, it would enable the receiving public to receive all services on the better standard, and in colour. Secondly, it will permit of the manufacture of single standard, 625 line only sets, which will be cheaper for the viewing public to buy than the dual standard set. Thirdly, concentration on the 625 line standard would improve export opportunity not only because this standard is in general use throughout Europe and widely elsewhere, but because it would widen the domestic basis for the growth of colour television at a time when overseas television services will be turning towards colour.

The Government statement agreed, of course, that the economies to be realized through better and cheaper sets has to be set against the cost to the two broadcasting organizations of duplicating the transmission of BBC 1 and independent television, but it was pointed out that this expenditure will be spread over a period of 10 to 15 years.

As outlined by the P.M.G. the present plan is that the duplicate services should start in London, the Midlands and the North simultaneously within the next three years. In this way, the duplicated programmes, including colour, could be available to nearly half the population from the outset with the expectation that some 75 per cent coverage would be achieved by the end of 1971.

When speaking in the House, the Postmaster General emphasized two important points. First, that there is no question of viewers' 405 line sets becoming useless overnight as BBC 1 and independent television will continue to be broadcast on 405 lines for as long as is reasonably necessary—which means a period of years. Second, it is of the essence that duplicate BBC 1 and duplicate independent television broadcasts on 625 lines should literally be duplicates of the 405 line services: they will not be BBC 3 and ITA 2. The whole point is to make the changeover from 405 lines to 625 lines and it is essential to the exercise that new, single standard sets shall be able to receive all television programmes.

The P.M.G. concluded his statement to Parliament by saying, "The House will, I believe, recognize that this decision resolves the biggest single technical issue outstanding in the broadcasting field. It opens the way to the achievement of the objective of transmitting all services on the better line-definition standard, and only on that standard. It broadens the basis on which colour television will be provided. It marks the beginning of the end of the continuing expense of a dual standard. And it will provide a firm base for long-term planning both by the broadcasting authorities and the manufacturing industry."

So there it is: a decision has been made and forward planning can commence. While not everyone will agree with the solution that has been adopted, the majority will be agreed that it is a very good thing to have brought to an end a period of uncertainty which the television industry could ill afford. Whether the Country as a whole can really afford a duplicated television service with colour on three channels is another matter and whether it will really increase our export potential is also open to debate.

There is no doubt that if colour is to get off the ground at all then it must be provided on more than one channel: if as at one time proposed, colour had been confined to BBC 2 on 625 lines it seems hardly likely that many colour sets would have been sold or rented, for not many people would have paid £250 to £300 or a rental of some 35s a week for two or three hours of colour a day on one channel—and the least popular channel of all, at that! Equally, a dual standard colour receiver is almost unthinkable, the complications of colour are great enough without that. So what other reasonable solution could there be? But it is not without its snags and, apart from the one of the immediate cost, the greatest of these may well prove to be the provision of adequate national coverage on u.h.f. Indeed the high percentage of population coverage which has been achieved on v.h.f. may never be duplicated on u.h.f., and to approach it may well require hundreds (or even thousands) of small relay stations.

Undoubtedly the plan is warmly welcomed by the two broadcasting authorities and certainly so by the capital equipment manufacturers who will be assured of a steady flow of orders from the BBC, ITA and the programme contractors for a number of years to come. For the domestic receiver manufacturers the immediate future is not so certain. Sales of receivers are likely to be at a very low level in the next year or two as most viewers will delay replacing their existing sets until the 625 line u.h.f. services are available to them and even then it is likely that more people will rent than buy, particularly if they decide to opt for colour.

However, whatever the pros and cons may be it is now too late for argument and we can but look forward to the new era of television with hopeful anticipation.

Simplified Information Theory and Data Transmission

(Part 1)

By A. E. Cawkell*, C.Eng., M.I.E.R.E., Sen.M.I.E.E.E.

Information theory is explained with the minimum of mathematics and with some short cuts in the interests of providing a readable article for those who are not deeply involved in the field. The information content of messages is discussed, leading to the transmission rate of signals, considered as discrete and continuous elements, through a noisy communication channel. The entropy concept is introduced separately. The article continues with sections on coding, trading of band width, power, and time, methods of error correction and detection, timing and synchronization, and systems of modulation. Examples of the theory applied are restricted to data transmission via telephone lines in view of current interest in the use of the public networks for this purpose. Several modern systems are described in some detail.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 276)

THIS article concerns the sending of messages in the form of a sequence of electrical signals, from a transmitter to a receiver along an electrically conductive channel.

In a basic communication system, symbols are selected from a collection by an operator at the source, are encoded into electrical signals, and are transmitted along the channel to the receiver. The receiving operator decodes the signals back to symbols for consideration.

What is conveyed to the receiving operator by the symbols depends on the *a priori* knowledge which he may have about the message. If he possesses a list of the symbols in the order in which they are to be sent, no information is conveyed. If, however, he has no *a priori* knowledge, and the symbols are completely independent of each other, an equal amount of information is conveyed by each symbol (assuming a perfect channel).

Usually the receiving operator's knowledge lies somewhere between these two extremes. Suppose he knows that the sending operator will only be sending signals which represent the letters of the alphabet. He can dismiss the possibility that the signals may occasionally represent numbers; this reduction in the range of possible symbols removes some doubt or uncertainty. If he also knows that the symbols are supposed to represent English words, his doubts about the symbols are further reduced. When he finally receives the message correctly, all his initial uncertainty is removed.

The information conveyed by the message is the amount by which his *a priori* doubt is lessened. The amount of reduction of doubt depends on the information that he possesses before reception.

'Information' in this context has a much more limited meaning than when the word is used in normal conversation. 'Data' as in 'data transmission', means signal sequences to which a meaning can be assigned; such signals may convey a very little or a great deal of information.

Redundancy in a message is a measure of the doubt which was not removed by receipt of the message. However, if redundancy is introduced at the source by transmitting more symbols than are strictly necessary, it is more likely that the message will be received correctly; for instance, the same message may be transmitted twice.

For the moment these statements will suffice; in information theory, as will be seen, meanings are further limited as one is concerned with the correct reception of messages, not their interpretation.

Information Content of the Discrete System

To maximize the amount of information conveyed (in an ideal situation unlikely to be encountered in practice), the occurrence of each symbol must be equally probable and there must be no *a priori* information at the receiver.

To fulfil these conditions, the sending operator would have to choose symbols from a collection at random. Suppose the collection consists of 16 symbols. The probability of receiving any symbol is $1/16^{th}$. The concern here is only with correct reception of a message, not its meaning. A convenient way of expressing the information content of each symbol is to count the minimum number of 'yes-no' decisions necessary to identify any one particular symbol from the collection. In this case, by successively halving, the selection would be from 16 to 8, 4, 2, 1, so that four 'bits' of information are necessary to identify any symbol. The information content is $H = \log_2 16 = 4$ bits. This concept also fits the case when there is only one symbol to choose from, as $H(1) = 0$. No information would be conveyed by successive reception of the same symbol.

Generalizing

Information content, $H = \log_2 n$ bits per symbol .... (1)

where n is the number of independent symbols from which the choice is made. The fact that n may not be a power of 2 does not preclude the use of this equation, because in information theory very long sequences of symbols are normally considered in order to determine average information content, when it may be shown that the equation is still valid.

The symbols may be simply encoded for electrical transmission using signals of the 'current-no current', 'mark-space', or '0, 1' type. If the 0, 1's represent elements of a binary number, four such elements would be required to identify any of the 16 symbols in the above example; for instance 1101 = 13. In this context the term 'bit' is used as an abbreviation for 'binary digit'.

For 16 symbols, there is always a probability of reception of $1/16^{th}$. If the probability of reception is not the same—as for instance when receiving letters which form

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words—the information content per symbol may be obtained only by long term averaging. Consider the average information content in bits per symbol for long messages with different probabilities per symbol—say as with e's and x's in the English language. The information content of any symbol i, occurring with a probability p_i, is $-\log_2 p_i$ bits. As the probability is always less than unity it is always negative, and $\sum_1 p_i = 1$.

The average value in bits for the information content $\overline{H}$ of independent symbols $a, b, \ldots,$ is $-(p_a \log_2 p_a + p_b \log_2 p_b + \ldots)$ so for the i^{th} symbol:

$$\overline{H} = \text{average of } \log_2 p_i = \sum_1 -p_i \log_2 p_i \text{ bits/symbol} \quad \ldots (2)$$

Suppose a long message is composed of symbols a, b, c, and d occurring with probabilities 1/2, 1/4, 1/8, and 1/8 respectively; then:

$$\overline{H} = -(1/2 \log_2 1/2 + 1/4 \log_2 1/4 + 1/8 \log_2 + 1/8 + 1/8 \log_2 1/8)$$

$$= 1/2 + 1/2 + 3/8 + 3/8$$

$$= 1\tfrac{3}{4} \text{ bits per symbol.}$$

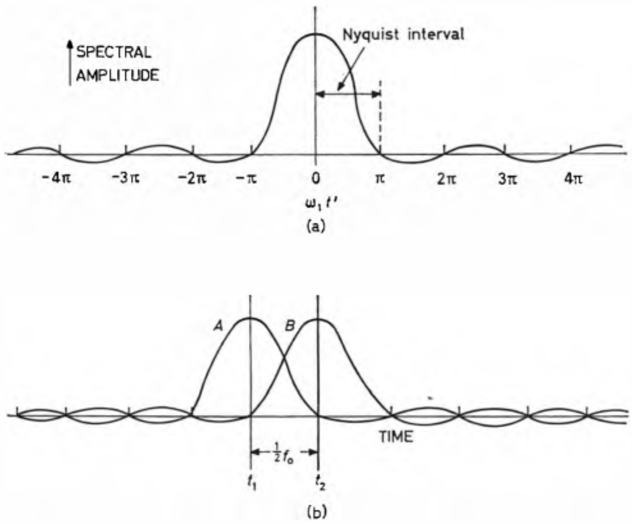

Fig. 1 (a). Time representation of spectral components, $\frac{\sin \omega_1 t'}{\omega t'}$ at the output terminals of an ideal filter, with an impulse input

(b). Spectrum at the output of an ideal filter with impulses spaced at $t' = 1/280$ applied to the input. At the instant t_1, no components of B interfere with A.

Information Rate of Discrete Signals: Channel Capacity

Transmission Rate of Signals Through a Noiseless Channel

If an impulse is applied to an ideal low-pass filter, that is to a filter with unity transmission response up to a frequency $2\pi f_0$ and zero transmission above that frequency, the amplitude/phase time relationships in the emerging waveform are of $\sin x/x$ form as shown in Fig. 1(a). The central peak and equally spaced axis crossings should be noted. If impulses are applied to the filter at intervals of $\pi/\omega_1 = 1/2f_0$, there will be no interference between output waveforms at the instant of peak as shown in Fig. 1(b).

Signalling may be carried out at a maximum rate of $2f_0$ impulses per second through a channel having the characteristics of an ideal filter cutting off at f_0Hz, whose bandwidth will be designated W Hz, so:

$$\text{Rate of transmission} = 2W \text{ impulses per second} \quad \ldots\ldots (3)$$

Critical adjustments would be necessary if the characteristics of a channel resembled such a filter, and no practical channel has such characteristics. However it may be shown that the equally spaced axis crossings, the absence of interpulse interference, and a proportionality between the area under the received waveform and the area under the trainsmitted waveform may be achieved, if the time representation of the waveform is of raised cosine form. This form may be achieved by a modification of the roll-off characteristics of the ideal filter with due regard for the shape of the input waveform. In practice, the fact that the waveform to be transmitted is not an ideal impulse and the channel is not an ideal filter does not inhibit signalling at a rate approaching the Nyquist rate because raised cosine waveforms are realizable in a practical situation.

Transmission Rate of Information Through a Noiseless Channel

The amplitude of the signal waveform was not discussed in the last section, but suppose there are a number of batteries connected in series with tapping points, and it is possible to alter the tap at the instant of making a mark or space. Signals would then be transmitted having peak amplitudes of $a, b, c, \ldots n$ volts according to the tap selected. These different voltages could be made to represent symbols, so if n amplitudes are available, the information content is $\log_2 n$ bits (from equation (1)) assuming that any symbol/signal state is equally probable. If the received waveform is of raised cosine shape, and is being received at the Nyquist rate, the information in successive waveforms is mutually independent only at one instant in time as has already been seen. This means that the receiver must know 'when to look' at the waveform, a question which is discussed later.

Combining these considerations with equation (3):

$$\text{Information rate, } H_r = 2W \log_2 n \text{ bits/second} \quad \ldots\ldots\ldots (4)$$

Transmission Rate of Information Through a Noisy Channel[1]

Noise of various kinds is always present in a communication channel, so that it is as if signals are being received from two sources, one generating information, and the other noise. The characteristics of the noise will vary from channel to channel, but for the purposes of information theory, noise amenable to definition is considered—Gaussian noise (noise impulses having r.m.s. values which occur with a probability given by the normal distribution curve).

Noise limits the number of levels, n, which can be received. For instance in a particular noisy channel it might be necessary to reduce the number of levels from 32 to 8 in order that the doubt, created by the noise, about which level is intended, is sufficiently small for the receiver to make an error only occasionally. In so doing $\log_2 n$ is reduced from 5 to 3—the penalty paid to ensure that signals are usually correctly received. The 'coarser' the level structure, the easier it is for the receiver to discriminate between levels. When one bit of information is signalled by switching on full power, and one bit is signalled by switching off, discrimination against noise is at a maximum.

The equation for the transmission rate through a noisy channel is:

$$H_r = H_{\text{sig+noise}} - H_{\text{noise}} \quad \ldots\ldots\ldots\ldots (5)$$

The derivation of this equation is outside the scope of this article, but it stems from the proposition that if two sources are generating sets of independent signals, the joint information content is the sum of the separate information contents:

$$H(x, y) = H(x) + H(y) \quad \ldots\ldots\ldots\ldots (6)$$

In equation (5) noise is regarded as a set of independent signals which are ' inaccessible '.

From equations (4) and (5):

$$C = 2W \log_{2n} (S + N) - 2W \log_{2n} N$$
$$= W \log_2 (1 + S/N) \text{ bits per second} \quad \ldots. \quad (7)$$

where C = channel information capacity
S = average signal power
N = average noise power.

It is necessary to qualify equation (7) by stating that at the rate C there will be an arbitrarily small number of errors as their occurrence is a matter of probabilities, also that a ' suitable form ' of encoding must be used. For a number of reasons the rate C has not, nor is likely to be achieved in practice, the most important reason being that an efficient enough encoding method is extremely difficult to devise. The criterion of equation (7) (the ' Shannon-Hartley ' law) remains a theoretical limit—an information rate which cannot be exceeded, but a target at which to aim, the nearness of approach being a measure of the efficiency of a communication system. It must also be emphasized that this discussion applies to statistically stationary signals—that is it is assumed that probability estimates made when observing very long messages will be the same no matter at what instant the observations commence. This assumption cannot be made in some forms of human communication; the example of the probability of occurrence of e's and x's in English, given previously, while illustrative of a point, is a bad example to the extent that the statistics of words may be a non-stationary process.

Comment on the Discrete Signal Concept

Gaussian noise was treated earlier as an inaccessible source of discrete signals, and information signals were considered as sequences of the states of a number of elements. This is a satisfactory concept as finite spacings are used which are amenable to definition and measurement. However, communication is often carried out using ' continuous ' waveforms; one can continue to use finite spacing concepts when discussing such waveforms, as will be seen, but often electrical signals having a continuous range of amplitude levels are analysed with the aid of integrals. This is not helpful as infinitesimally small spacing concepts are of little assistance in practice. As this approach is so often used in textbooks it will be included in the second half of the next section for the sake of completeness.

The discrete system, or consideration of a continuous system by sampling and quantization (to be described) involves the use of binary information, as indicated by the prefix ' $\log_2$ '.

When using integrals, logarithms to the base e are inherent in the mathematics.

The conversion factor is:

1 natural unit = 1·443 bits

The ' Continuous ' System

It may be shown from Fourier analysis that a periodic time function $f(t)$ containing frequency components up to W Hz may be completely reconstructed by specifying its ordinates at a series of points t_1, t_2, ..., spaced at intervals of $1/2\,W$ seconds.

The amplitude of any ordinate may be quantized into very small increments Δx. In an interval of T seconds, $2WT$ samples of n levels comprised of $\Delta x n$ increments will completely specify the function. Sampling more frequently will not provide any additional information about the waveform as there are only n independent ordinates when $n = 2WT$.

The considerations are similar to those decribed previously. A continuous waveform is shown in Fig. 2(a), with the requisite sampling points t_1, t_2, ..., at intervals of $1/2\,W$ seconds, shown in Fig. 2(b). The sampling impulses emerge from the ideal filter with a $\sin x/x$ form at intervals of $1/2W$, amplitude modulated by the information waveform (Fig. 2(c)). The coincidence of zeros will be observed. The sum of all the components of the successively received waveforms is a reconstruction of the original waveform. In practice the actual filter is a channel whose characteristics have been adjusted to resemble a filter from which raised cosine waveforms will emerge with due regard for the shape of the input waveform. The received waveform is now being treated in its entirety, with a departure from the idea that the information may be recovered by synchronous sampling at the receiver already discussed previously, so it is necessary to consider the information content of a continuous system, as was done for the discrete system.

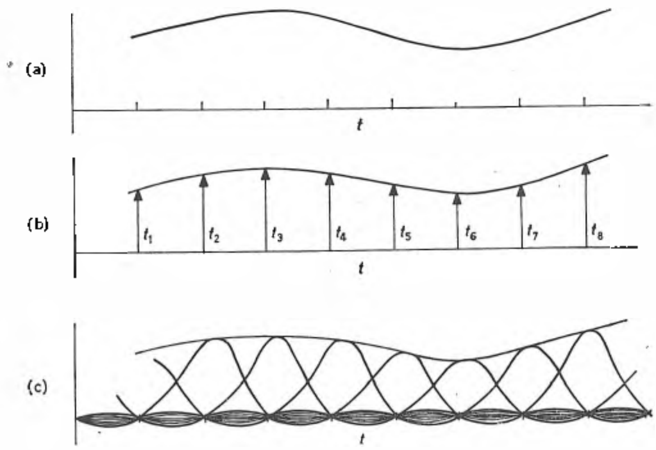

Fig. 2 (a). Original waveform, band limited to W Hz

(b). Sampling impulses at t_1, t_L .... where $\Delta t = \frac{1}{2}W$sec

(c). Reconstruction of original waveform by passing the sampling (modulated) impulses through an ideal filter of bandwidth W Hz, with reconstruction by the components of the emerging $\sin x/x$ waveforms

This may be derived from equation (2) by arranging that p_i is replaced by the probability density $p(x)$ associated with a message source $x(t)$:

$$\overline{H}(x) = \int_{-\infty}^{\infty} p(x) \ln p(x)\, dx \quad \ldots\ldots\ldots. \quad (8)$$

The summation sign has been replaced by the integral, and the differential interval Δx has been reduced to the infinitesimally small interval dx of the probability density $p(x)$.

This is a mathematical representation of the average information content of an electrical signal having a continuous range of amplitude levels with an infinite information content—an unrealizable condition; however, constraints will be introduced.

The electrical form of signals constrained by power which theoretically can convey the maximum amount of information is the Gaussian distribution; this may be regarded as the most efficient way of encoding signals with the power available, as the distribution density of the signal states (amplitudes) has the smallest variance from the average value of the distribution density; the maximum possible 'disorder' exists, so that provided the system can be ordered it provides the opportunity for the highest possible information content. The curve for this probability density function is described by:

$$p(x) = 1/2\pi\sigma \exp(-x^2/2\sigma^2) \quad \ldots\ldots\ldots (9)$$

where σ^2 = mean power.

This curve also describes, of course, the amplitude probability distribution of Gaussian noise.

The information content can now be found of the continuous system constrained by power, using it in the most efficient manner, and accompanied by Gaussian noise.

Re-arranging equation (9):

$$-\log p(x) = \ln \sqrt{(2\pi\sigma^2)} + \ln[\exp(x^2/2\sigma^2)]$$

and substituting in equation (8):

$$\overline{H}(x) = \int_{-\infty}^{\infty} p(x) \ln \sqrt{(2\pi\sigma^2)}\, dx + \int_{-\infty}^{\infty} p(x) \ln[\exp(x^2/2\sigma^2)] dx$$

which gives:

$$\overline{H}(x) = \ln \sqrt{(2\pi e\sigma^2)} \text{ natural units} \quad \ldots\ldots (10)$$

It has already been seen that information may be trans-

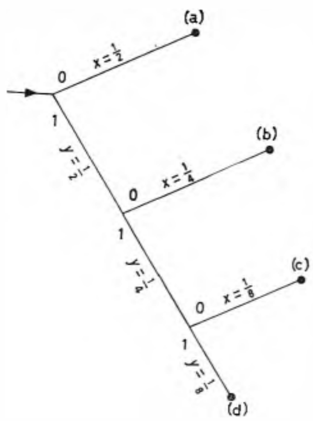

Fig. 3. Construction of a code

mitted at a rate of $2W$ elements per second (equation (4)), so combining this with equation (10):

$$H_r = 2W \ln \sqrt{(2\pi e\sigma^2)}$$

$$= W \ln \sqrt{(2\pi e\sigma^2)} \text{ natural units per second} \quad \ldots\ldots (11)$$

which is the information rate in a continuous system through a noiseless channel limited by bandwidth and power.

The transmission rate through a noisy channel was given by equation (5), so substituting equation (11) in (5), but using S and N to represent signal and noise power:

$$C = [W \ln 2\pi e(S + N)] - W \ln 2\pi eN$$

$$= W \ln (1 + S/N) \text{ natural units per second} \quad \ldots. (12)$$

Which is the theoretical channel capacity for the continuous system (compare with equation (7); the remarks which followed that equation also apply here).

Entropy

The word entropy, borrowed from thermodynamics, is often used in information theory because of a certain resemblance between an ensemble of states in a message source, and the states of particles in a gas; in both cases the states are described by probability functions.

In information theory, entropy has been described as a 'measure of disorder', the idea being that a system having equi-probable states is in a state of maximum entropy.

Equation (8) with the sign reversed (entropy being the negative of information) has been represented as the entropy of a message source, being that state of the source representing the maximum doubt at the receiver—the receipt of any symbol/signal is equally probable. As information is received so is entropy reduced at the same rate. A system in a state of maximum entropy is available for ordering into a system having maximum information content, assuming that this process of ordering is possible.

Noise is often in a state of maximum entropy, but it is not accessible for purposeful ordering. The Gaussian distribution is the state of maximum entropy for noise or signals limited by power.

Filtering

The amount by which the channel capacity falls short of the Shannon ideal in a practical system may be lessened by using carefully designed filters. To recapitulate, the object of filtering is to ensure that there is no interaction between successive waveforms, that the possibility of distinguishing signals from noise is maximized, and that the signal waveforms are of a shape to permit signalling rates to approach the Nyquist rate.

The optimum signal shape is of modified $\sin x/x$ form —the 'raised cosine' form, so called because of its spectral amplitude functions. The transmitting filter is designed so that when considered with the channel characteristics and the receiving filter (which is designed primarily for signal-to-noise improvement) the raised cosine shape and minimum inter-symbol interference is achieved. Design procedures are described in the literature.

Coding

STATISTICAL MATCHING

The Morse code was an early method of encoding symbols into electrical signals. Instead of allocating combinations of long and short 'marks' to each symbol without consideration, short combinations were allocated to letters occurring frequently in the English language (e = -), and the longer combinations were allocated to less frequently occurring letters (Q = — — - —). As a result, the transmission of a message in English may be accomplished more quickly than would otherwise be the case.

A similar approach may be carried out for any message where the probability of occurrence of symbols has been established. Suppose a message is composed of combinations of only four letters which occur with the probabilities—

$$a = 1/2$$
$$b = 1/4$$
$$c = 1/8$$
$$d = 1/8$$

and that these letters are generated by a message source at the rate of one per second. There are four possible symbols, requiring $\log_2 n = 2$ bits per symbol for transmission. If transmission is to take place at the same rate as generation, the channel capacity must be 2 bits per second. The average information content is $1\frac{3}{4}$ bits per symbol as was shown earlier. In order to construct the most efficient code the following procedure may be carried out (see Fig. 3)[2].

Draw a line diagram so that at each branch the lines x, y, terminate in a symbol or group of symbols which have equal or nearly equal probabilities. Label the bifurcation 0, 1. Determine the code group to be used for each symbol by reading off the 0, 1's on a line from the origin to the symbol.

From Fig. 3:

$$a = 0 \quad = 1 \text{ bit}$$
$$b = 1\,0 \quad = 2 \text{ bits}$$
$$c = 1\,1\,0 = 3 \text{ bits}$$
$$d = 1\,1\,1 = 3 \text{ bits}$$

The average number of bits per symbol in a message is the average of their occurrence:

$$\overline{H} = (1/2 \times 1) + (1/4 \times 2) + (1/8 \times 3) + (1/8 \times 3)$$
$$= 1\tfrac{3}{4} \text{ bits.}$$

The channel capacity required to transmit information at the rate of generation is now $1\frac{3}{4}$ bits/second, compared to the original requirement of 2 bits/second, a 12·5 per cent reduction, and the efficiency is 100 per cent as the average information content per symbol equals the channel capacity—source and channel are statistically matched.

The above example is convenient because of the ease of division into equal probabilities; improvements can be obtained if the branches lead to symbols or groups of

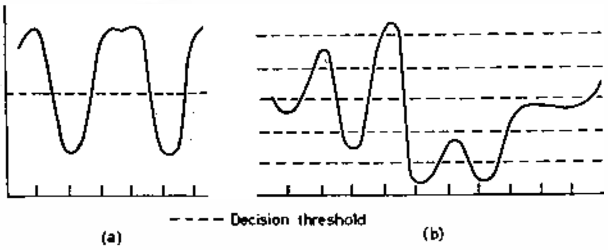

Fig. 4 (a). Two level signal
(b). Six level signal

symbols with approximately equal probabilities, but not 100 per cent efficiency.

One case where it might be expected that the procedure could lead to an improvement in efficiency is the transmission of English words. Theoretically improvements could be obtained, but the complexity of the coding system would probably offset the benefits.

Trading of Transmission Parameters

The Shannon-Hartley equation (7 or 12) defines the terms of trade between bandwidth, power, and time; intertrading can be used to suit a particular situation. For instance, an increase in bandwidth will result in a greater increase in channel capacity than an increase in signal power; bandwidth is a linear function of C, whereas signal power is a logarithmic function. If signal states are being transmitted at the limiting rate $2W$, the possibility exists of increasing the information rate by increasing $\log_2 n$, the number of elements per state, depending upon the channel noise and the acceptable error rate.

The information rate possible through a particular channel depends on the system application. In a digital signalling system, say for the remote operation of a computer via a telephone line, an error rate of not less than 1 in 10^7 may be required. If the same channel were to be used for signalling the output voltage from a number of pressure transducers with a 1 per cent accuracy requirement, a much faster rate of transmission is obviously possible.

The possibility of increasing $\log_2 n$ in the presence of noise is illustrated in Fig. 4. For a two level signal (Fig. 4(a)) one threshold is established at the receiver, and for six levels, five thresholds must be established. For the same signal power, the margin for error with noise present is on average five times less for the six level than for the two level signal; generalizing, an n level signal has a margin of error $1/(n-1)$ times that of a two level signal.

Suppose conditions will permit the transmission of an eight level signal, but bandwidth is at a premium. A method of eight level encoding is shown in Fig. 5. The binary elements are taken three at a time and are transmitted as any one out of eight levels, using the code shown. The information rate in Fig. 5(b) is the same as that in Fig. 5(a) but the channel bandwidth requirements have been reduced three times at a penalty of $1/8-1 = 1/7^{\text{th}}$ the margin of error against noise (accept-

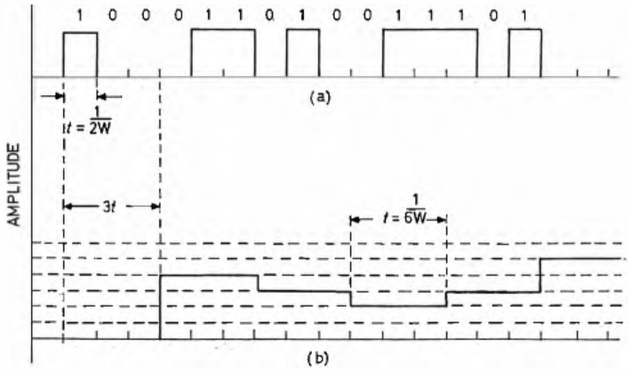

Fig. 5. Recoding a two level (two state) signal into an eight level signal
(a) Binary *(b) Eight level*

Code

Binary	*No. of levels*
0 0 0	0
0 0 1	1
0 1 0	2
0 1 1	3
1 0 0	4
1 0 1	5
1 1 0	6
1 1 1	7

able in this case). It will be noted that the re-encoding gives rise to a delay of $3t$, which is the time taken by the re-encoding device waiting for the first three bits of information, the transmitted information being $3t$ 'late' thereafter.

In contrast, suppose the importance of a low error rate is paramount, and it is desired to re-encode multi-level or analogue information into two state form in order to obtain the best discrimination against noise. If a continuous waveform is sampled at intervals of $\frac{1}{2}W$ seconds and quantized into n levels, the information capacity of the channel must be $2W \log_2 n$ bits/second. Now each ordinate could be signalled by suitable re-encoding as m pulses of s levels per pulse, that is $\log_2 n = m \log_2 s$; the channel capacity must now be $2W\, m \log_2 s$. The information rate is unaltered so that:

$$2W \log_2 n = 2W\, m \log_2 s$$
$$m = \log_2 n / \log_2 s$$

and for a two level signal:

$$m = \log_2 n$$

In other words re-encoding into a two level signal has increased the bandwidth by $\log_2 n$ times the bandwidth required for the multi-level signal. This system is known as pulse code modulation (p.c.m.). The system brings with it the option of using polar signalling and reduced power as described later in this article; moreover with two level signalling a relatively small increase in signalling power brings with it a much reduced error rate.

This method has been used to transmit speech waveforms quantized into 32 levels (Fig. 6). A disadvantage is the introduction of quantizing noise; random errors occur during the process of deriving finite steps from the continuous signal, and the finer the steps the lower the noise. With speech, 32 level quantization reduces this noise to a fairly low level. If the highest speech frequency to be transmitted is 10 000Hz, the transmission rate for 32 levels is $2W \log_2 n = 10^5$ bits/second. Re-encoding by p.c.m., the bandwidth multiplying factor is $m = \log_2 n = 5$, so a bandwidth of 50 000Hz is required.

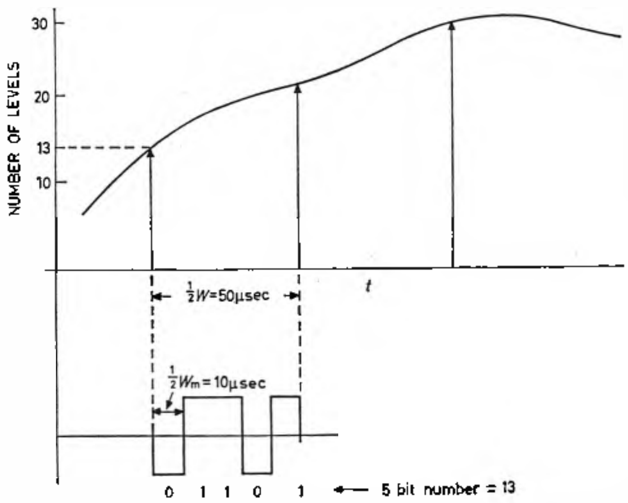

Fig. 6. Loding an analogue signal (W = 10KHz) into a two level signal (p.c.m.)

Error Detection

The ideal form of coding necessary to achieve the theoretically possible channel capacity could be attained by using very long sequences of equi-probable signals to represent symbols—say by signalling in binary code, and choosing the sequences of 0's and 1's by tossing a coin. The receiving operator or decoding device would be required to consult a code book or store, holding all the possible sequences, and would have to compare each received sequence with all the possible sequences to decide which sequence was intended. The longer the sequences, the longer the time taken to make the decision, but the more likely would it be that the decision would be correct. The decision could be made with a relatively poor signal-to-noise ratio.

This procedure would involve prohibitive delays and equipment complexity; it has been found in practice that the best compromise is reached if signalling is carried out at a fast rate, accepting the fact that there will be errors; error detection and/or correction methods are used to achieve the desired net error rate at a penalty of some delay and equipment complexity.

One detection method which is often used is to add redundant elements to the message. Two examples known as 'cyclic' codes, and two other examples of the many arrangements which have been developed utilizing redundant elements, will be given to illustrate the principles.

Suppose a message is transmitted in blocks of six bits as shown in Fig. 7(a). A 'parity' bit is added at the end of each block according to the rule that the total number of 1's in the block must be odd. Any single error in the block may be detected if the receiver knows the rule.

To carry the process further with better protection, consider Fig. 7(b). Here the detection of single or double adjacent errors is possible, as the rule is that the total of the 1st, 3rd, and 5th bits and the first parity bit must be odd, and the total of the 2nd, 4th, and 6th bits and the second parity bit must be odd. For instance if the first block in the figure is received as 11011000 there must be two errors, although if there were four errors—say 01011100—detection is not possible.

Once an error has been detected, a request for a re-transmission of the block may be sent.

The two co-ordinate parity bit code is efficient as redundancy is low, and is effective against Gaussian and bursts of impulsive noise; it is therefore particularly suitable for information signalled over telephone lines. A matrix consists of m rows each of k information bits (Fig. 7(c)). Parity bits are added by reference to both horizontal and vertical rows. With a matrix consisting of 32 6-information bit plus 1 parity bit rows, and 1 parity bit per column, there are 39 parity bits out of a total of 321 bits—17 per cent redundancy. Single, double and error bursts up to k bits long may be detected.

6 bit block	*Parity*	*6 bit block*	*Parity*		*k bits* →	*Parity*
1 0 1 1 1 0	1	1 0 1 1 1 0	0 0	↓	1 0 1 1 0 1	1
1 1 0 0 1 0	0	1 1 0 0 1 0	1 0	m	1 1 1 0 1 1	0
0 0 0 0 0 1	0	0 0 0 0 0 1	1 0	bits	0 0 0 1 1 1	0
0 1 0 1 1 1	1	0 1 0 1 1 1	0 0			
			Parity		1 0 1 1 0 1	1
(a) Single error detection: 16⅔% redundancy		(b) Single and double adjacent error detection: 33⅓% redundancy			(c) Co-ordinate code single, double, and error burst detection up to k bits long (if $k = 7$ (incl. P), and $m = 33$ (incl. P), redundancy= 17%)	

Fig. 7. Error detection by parity bit check

In yet another form of coding known as 'fixed weight' codes, p out of q possible combinations of elements are assigned meanings in such a way that an error is likely to produce a q combination. The number of sequences to be assigned out of 2^q sequences may be found from $q!p!\ (q-p)!$, so a '4 out of 8' code produces $8!/4!\ (8-4)! = 70$ sequences, and 70 out of a possible 2^8 sequences are made meaningful, which provides 70 sequences which contain four identical elements—for instance exactly four 1's. Any odd number of errors and many even errors may be detected.

This type of code has been used[3] for detection with automatic request for repetition. The system consists of a duplex channel in which A transmits continuously to B and vice versa. A telegraph code is recoded into a 3 out of 7 code, and the detection of an error by say, B, initiates a transmission from B to A automatically, interrupting A's transmission, and asking for a re-transmission of the last few signals. The whole cycle of detection to re-reception lasts for the duration of a few signals only.

Error Correction

Methods of error correction usually depend on the use of words (blocks of symbols) which differ from any other words by at least three symbols. This difference is known as the 'Hamming distance' after the originator of the scheme[4]. If an error occurs, the erroneous word still resembles the intended word more closely than any other word. If the Hamming distance is increased to five, two errors may be detected and corrected[5]. The penalty is considerable redundancy. An example of a word in this system is shown below.

Position of symbol	1	2	3	4	5	6	7	8
Significance of symbol; (parity or power of 2)	P	P	2^3	P	2^2	2^1	2^0	P
Word example	1	0	1	1	0	1	0	0

The example is decimal 10 in binary, the elements being placed in positions 3, 5, 6 and 7. Parity bits in positions 1, 2, and 4 are used for single error detection, and in position 8 for double error detection. The redundancy is 50 per cent. The rules are as follows:

Check A; sum of elements in positions 1,3,5,7, must be even

Check B; sum of elements in positions 2,3,6,7, must be even

Check C; sum of elements in positions 4,5,6,7, must be even

Check D; sum of all elements must be even.

If a single error is made, check D will show odd parity.

If checks C, B, and A are now made in that order, and a note is made as a 0 for even and a 1 for odd parity, the binary number so formed indicates the exact position of the erroneous element which may be then corrected.

For instance if 10110100 is received as 00110100, check D shows odd parity so there is an error. Check C even = 0, check B even = 0, check A odd = 1. The number 001 indicates that the error is in the first position.

Moderately involved circuits are required for detection and correction. Shift registers and combination logic circuits are used. A shift register is a form of store consisting of a cascade of bistables each storing a 0 or a 1, controlled by an input pulse to the first stage, and by timing pulses fed to each stage. Blocks are assembled ready for transmission with information and parity bits in the correct positions. At the receiver each block is stored sufficiently long for identification and correction to be carried out in register/logic circuits with feedback loops.

Applications of Error Detection and Correction Systems

The economics of a system consisting of error detecting and/or correcting arrangements is related to the purpose of the channel and the cause of errors. Errors may be caused by Gaussian noise, by single impulses which destroy one or several elements, or by noise bursts which destroy many elements. One approach is to use a specially designed error correcting system to correct the type of error which occurs most often, backed up by detection which is needed when a correct signal is interpreted as an error, or when infrequent errors occur of a kind not catered for in the error correction system.

Usually economic considerations result in a choice of one of the many efficient detection codes, with a 'request for repeat' system.

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(*To be continued*)

Goonhilly to be Extended

The Post Office has placed an order with the Marconi Co. Ltd for the major components of a new satellite earth station terminal for use with communication satellites. The whole installation is to be provided, by April 1968, at the existing P.O. earth station at Goonhilly, Cornwall.

Among the equipment to be provided by Marconi is an aerial 90ft in diameter which will direct radio beams to a new multiple access satellite to be positioned over the Atlantic in 1968. The new satellite, to be called Intelsat III, will be of the stationary type, like Early Bird, only considerably larger. It will enable simultaneous telephone communication to be set up with countries in North and South America, the West West Indies and Africa. Television programmes can also be exchanged.

The aerial, a limited steerability model based on the Post Office Consultant's design, will cover a 210° arc in azimuth, to enable it to work both the Atlantic and Indian Ocean satellites. The entire 550ton structure, including transmitters, receivers and ancillary equipment, will pivot in azimuth, driven by four s.c.r.-fed d.c. motors generating a total of 60 h.p.; these will be able to point the entire structure with an accuracy better than one minute of arc in winds of 47 mile/h gusting to 66 mile/h when the autotract facility is in operation. Additionally, the aerial can be moved from one satellite to another at a speed of 10°/min in azimuth and 5°/min in elevation.

The transmitter will use wideband travelling wave tubes to provide the final peak saturation power output of 10kW. A t.w.t. was chosen for its overall system performance in a wideband system in preference to klystrons since each individual carrier would require a complete klystron transmitter whereas the entire 500MHz band can be covered by a simple t.w.t. amplifier. The t.w.t. is also likely to prove more reliable, as no mechanical tuning is involved. The gain of each of the t.w.t. stages to be used will be of the order of 30dB, throughout the entire civil satellite communications band of 5 925MHz to 6 425MHz.

The transmitting facilities available will be adequate for over 500 telephone channels and one television channel simultaneously, using a multi-access satellite of the Intelsat III type.

Solid-state circuits are used throughout the transmitting system, with the exception of the two travelling-wave-tube amplifiers.

The receiving system will be based on the use of a low noise parametric amplifier mounted on the back of the dish structure, to provide a combination of low noise performance with a very wide bandwidth of 500MHz.

A closed-cycle cryogenic system will use gaseous helium operating at approximately −257°C, close to the liquid helium temperature. This system has the advantage that it will operate for long periods without any need for topping up.

The parametric amplifier will consist of three identical gallium arsenide varactor diode stages, connected in cascade. These three stages will be mounted, together with their associated circuits, inside the low temperature enclosure. Each of these three stages will be fed from a klystron pump source through a three-way passive splitter. These klystrons will be the only part of the receiving system which does not use completely solid state equipment. It will provide 30mW of pump power at a frequency of 34GHz.

A low noise tunnel diode amplifier will form the second main amplifying stage in the receiver, and this will also be mounted on the back of the aerial structure, in the same package as the parametric amplifiers. The tunnel diodes will operate from the same low voltage supplies as the parametric amplifiers, and will require no routine maintenance.

The receiver will cover the complete satellite communications band from 3 700MHz to 4 200MHz, including all possible channels from both the Intelsat II and III type satellites, as well as from the Intelsat I (Early Bird) satellite.

A Simple Active Filter with Independent Control over the Pole and Zero Locations*

By W. Farrer†, M.Sc.

The article describes an active circuit arrangement using a modified parallel-T network, an amplifier with unity positive gain and two potentiometers such that either low or high-pass filter characteristics may be arranged. The transfer function in each case consists of two poles and two conjugate zeroes. Variation of one of the potentiometers sets the location of the zero on the frequency axis while the other potentiometer adjusts the position of the poles so as to be able to obtain a maximally-flat response in the pass-band.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 276)

A SCHEMATIC of the active filter is shown in Fig. 1, where the network between points X_1, X_2, Y and Z is a normal parallel-T network modified such that the resistive series arm $X_1 Y$ is fed with signal V_1 volts while the capacitive series arm is fed with KV_1, where $0 \geqslant K \geqslant 1$. The common end of the series arms of the parallel-T, point Y, is connected to the output (V_o) via an amplifier of unity positive gain, high input impedance and low output impedance; such as an emitter or cathode follower. A fraction m of the output is fed back and applied to the common end of the shunt arms of the parallel-T, point Z, where $0 \geqslant m \geqslant 1$. Both KV_1 and mV_o may be derived from low impedance potentiometers across V_1 and V_o respectively.

The condition $K = 1$, $m = 0$, corresponds to a conventional parallel-T and the circuit arrangement then has the transfer function:

$$V_o/V_1 = \frac{1 + p^2 T_3 T}{p^2 T T_3 + p[T(1+R_1/R_2)+T_3(1+C_3/C_2)] + 1} \quad \ldots\ldots\ldots\ldots (1)$$

where:

$$T = R_1R_2C_1/(R_1+R_2) = R_3(C_2+C_3) \quad \ldots\ldots\ldots\ldots (2)$$

$$T_3 = (R_1+R_2)C_2C_3/(C_2+C_3) \quad \ldots\ldots\ldots (3)$$

and the operator $p = d/dt$.

Equation (2) is the necessary and sufficient condition to ensure that the parallel-T network connecting points X, Y and Z shall have a true zero in its transfer function.

An analysis for other values of K and m yields a transfer function given by:

$$V_o/V_1 = [1+Kp^2TT_3] \div \{p^2TT_3 + p[T_3(1+C_3/C_2) + T(1+R_1/R_2)](1-m)+1\} \quad \ldots\ldots\ldots (4)$$

If V_1 and KV_1 are interchanged at X_1 and X_2 respectively the transfer function becomes:

$$V_o/V_1 = K + p^2TT_3 \div p^2TT_3 + p[T_3(1+C_3/C_2) + T(1+R_1/R_2)]\,(1-m)+1 \quad \ldots\ldots\ldots (5)$$

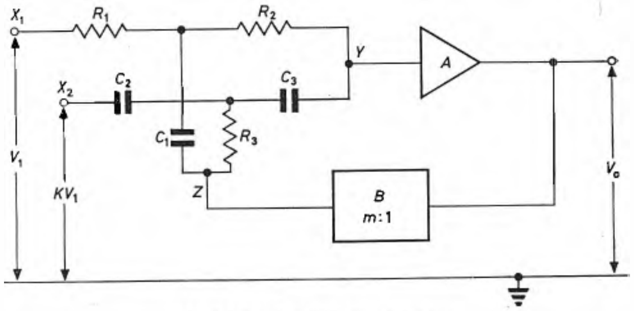


Fig. 1. Basic filter arrangement

For symmetrical parallel-T networks $C_2 = C_3 = C_1/2 = C$; and $R_1 = R_2 = 2R_3 = R$, giving $T_3 = T = RC$ and thus reducing the transfer function to:

$$V_o/V_1 = \frac{1 + Kp^2T^2}{p^2T^2 + p4T\,(1-m) + 1} \quad \ldots\ldots\ldots (4a)$$

and:

$$V_o/V_1 = \frac{K + p^2T^2}{p^2T^2 + p4T\,(1-m) + 1} \quad \ldots\ldots\ldots (5a)$$

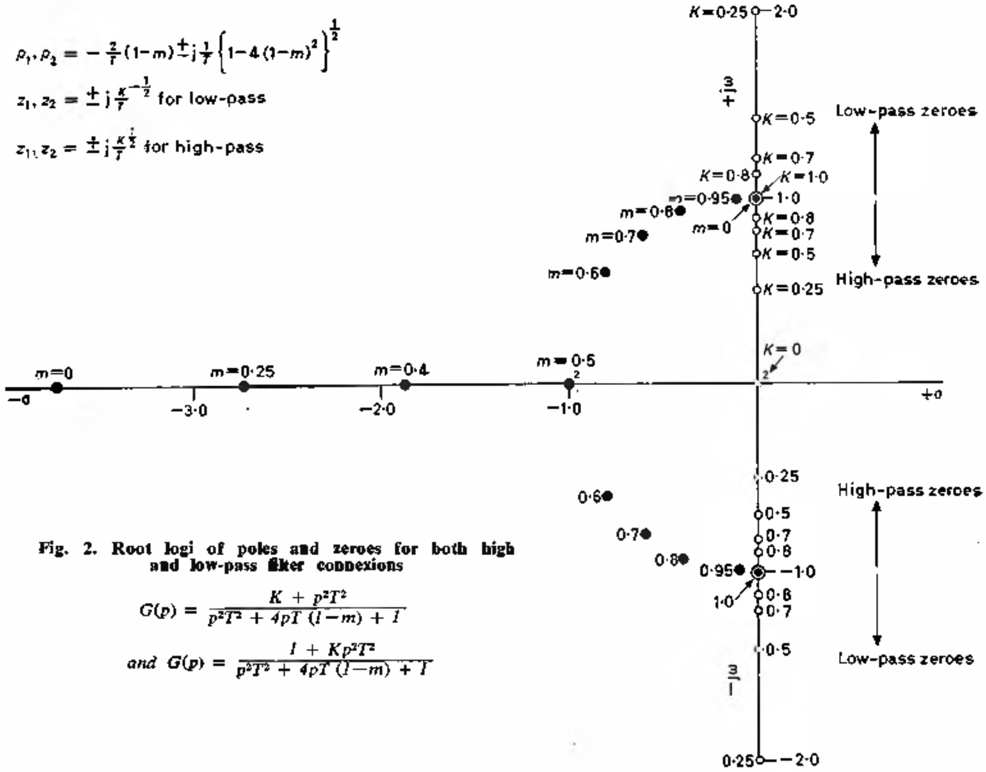


Fig. 2. Root loci of poles and zeroes for both high and low-pass filter connexions

$$G(p) = \frac{K + p^2T^2}{p^2T^2 + 4pT\,(1-m) + 1}$$

$$\text{and } G(p) = \frac{1 + Kp^2T^2}{p^2T^2 + 4pT\,(1-m) + 1}$$

* *Provisional Patent.*

† *The Harland Engineering Co. Ltd.*

The input connexion giving equation (4a) for the transfer function will be termed the low-pass connexion while the connexion giving equation (5a) for the transfer function will be termed the high-pass connexion.

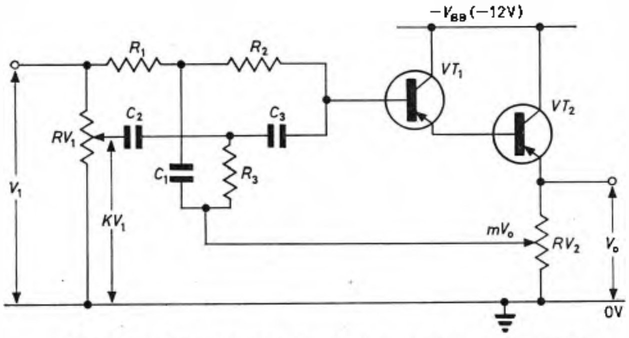


Fig. 3. A practical realization of the system shown in Fig. 1
$R_1 = R_2 = 2R_3 = R = 15k\Omega$; $RV_1 = 1k\Omega$
$C_2 = C_3 = C_1/2 = C = 0{\cdot}01\mu F$; $RV_2 = 500\Omega$
$\omega_0 = 1/RC = 2\pi \,.\, 1050$ rad/sec

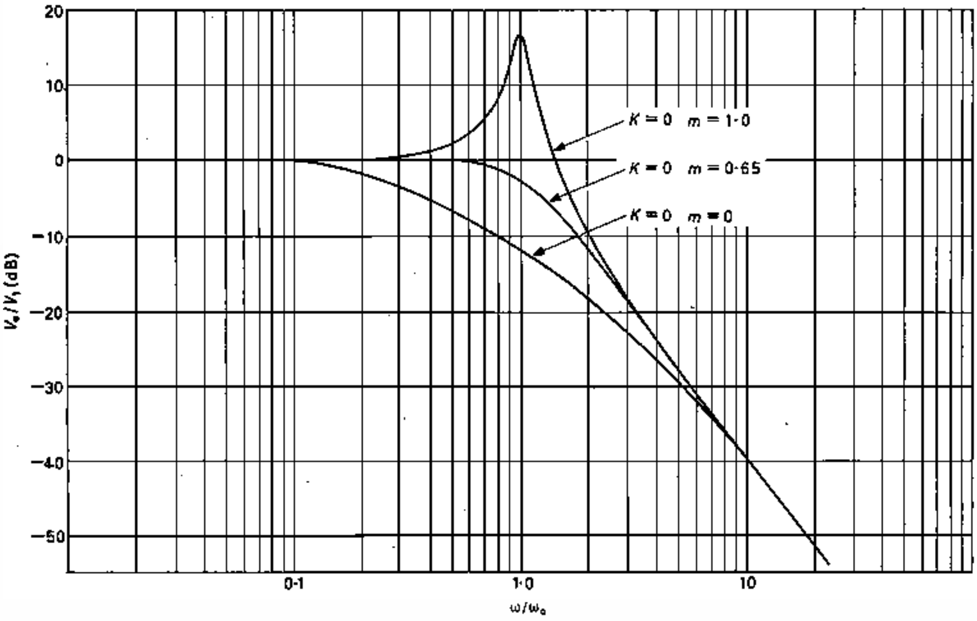
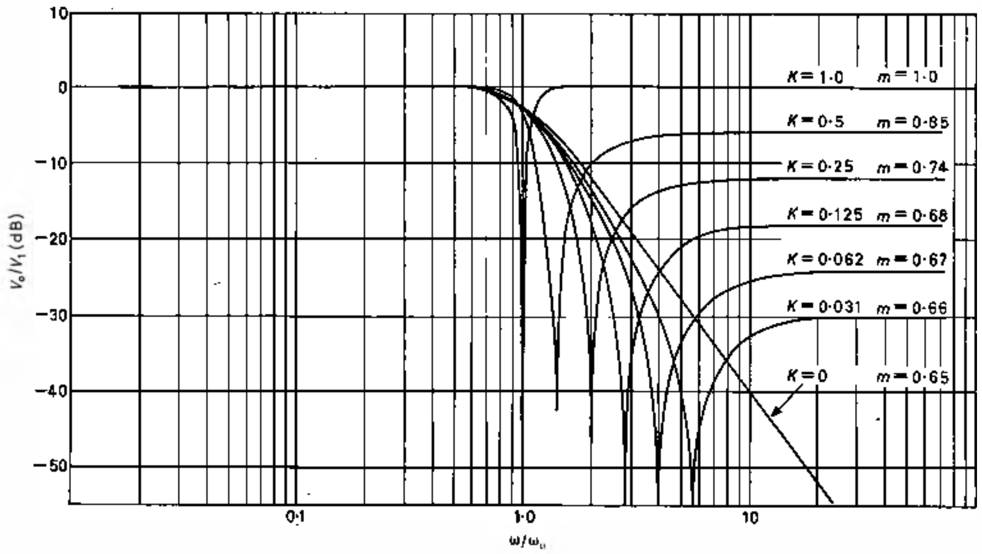


Fig. 4. Low-pass filter at $K = 0$ and various m

Fig. 5. Low-pass filter for various values of K with m adjusted to give −3dB at ω_0



The transfer function contains two conjugate zeroes Z_1 and Z_2 and two poles P_1 and P_2. The positions of the zeroes in the complex plane are given by:

$$Z_1,\ Z_2 = \pm jK^{0.5}/T = \pm j\omega_0 K^{0.5} \quad \ldots\ldots\ldots\ (6)$$

for the high-pass case, and:

$$Z_1,\ Z_2 = \pm jK^{-0.5}/T = \pm jK^{-0.5}\omega_0 \quad \ldots\ldots\ (7)$$

for the low-pass case.

Where:

$$\omega_0 = 1/T \quad \ldots\ldots\ldots\ldots\ldots\ (8)$$

The positions of the poles on the complex plane are given by:

$$P_1, P_2 = -2\omega_0(1-m) \pm j\omega_0[1 - 4(1-m)^2]^{-0.5} \quad \ldots\ (9)$$

for both high and low-pass connexions. It is seen that the poles are complex only for values of m greater than 0·5.

Fig. 2 shows a plot of the root-loci of the poles and zeroes for various values of m and K. Both high and low-pass connexions are represented. It is seen that the loci of the poles follows the arc of a circle of radius $\omega_0 = 1/T$ for all values of m from 0·5 to unity inclusive; while the loci of the zeroes is always along the $j\omega$ axis. The control exercised by m allows the poles to be located in the most advantageous position to achieve maximum flatness in the pass-band, while variation of K allows the frequency at which maximum rejection is required to be selected. The filter cut-off ω_0 is determined by the RC values of the parallel-T network and is constant for changes of m and K.

A practical realization of the proposed system is shown in Fig. 3 and uses transistors as the active elements. Two transistors are used in a 'Darlington' connexion to achieve the unity gain amplifier with high input impedance and low output impedance. Fig. 4 shows the variation of V_o/V_1 with frequency for various values of m at a constant K value of zero. Fig. 5 shows a family of curves of gain against frequency for various values of K, for each value of K the feedback m was adjusted so that the response was maximally flat and passed through the −3dB point at ω_0.

With X_1 and X_2 interchanged, and attention given to maintaining the correct bias drive to the transistors, the high-pass results were obtained and are shown in Fig. 6 and 7 respectively. Fig. 6 for

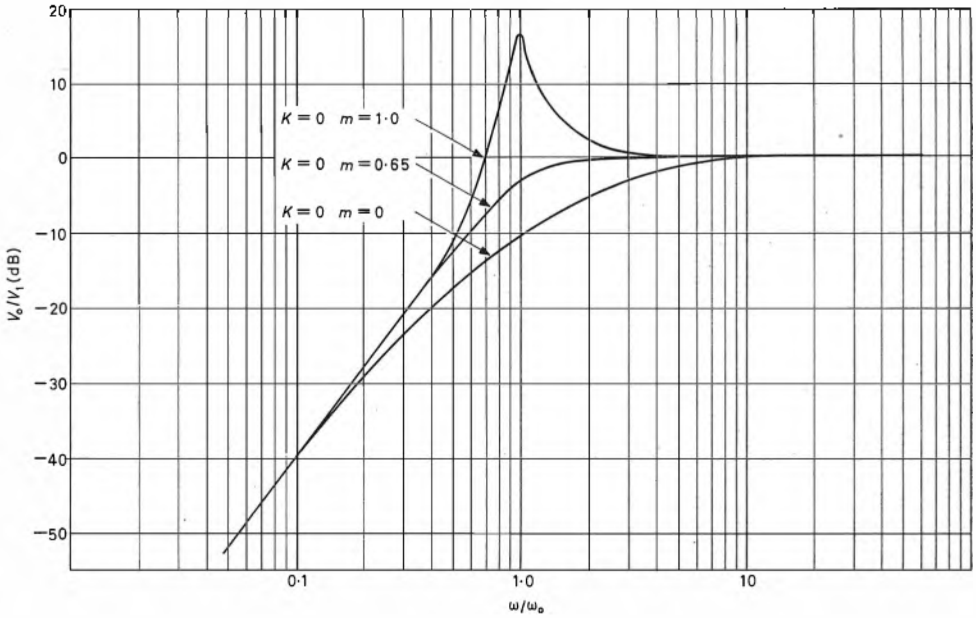


Fig. 6. High-pass filter for various m, $K = 0$

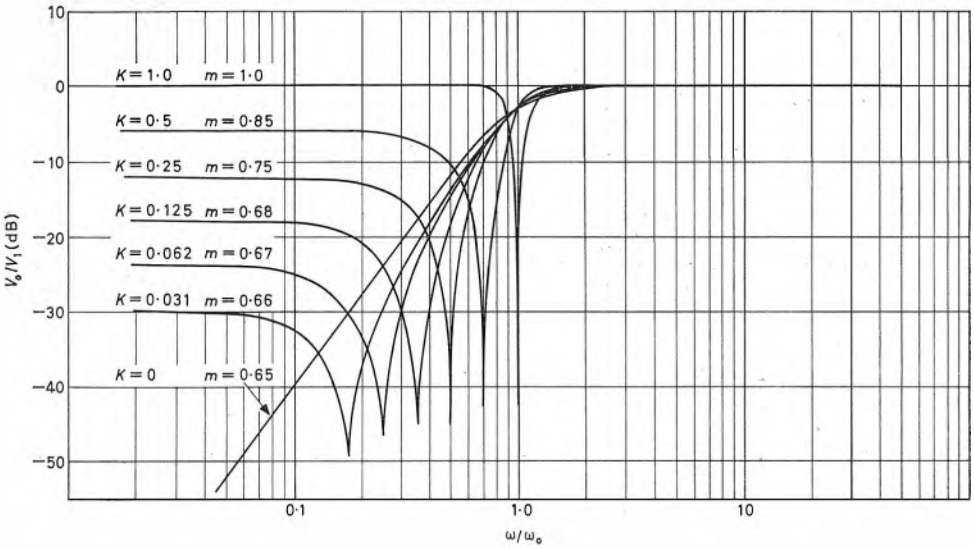


Fig. 7. High-pass filter for various values of K with m adjusted to give −3dB at ω_0

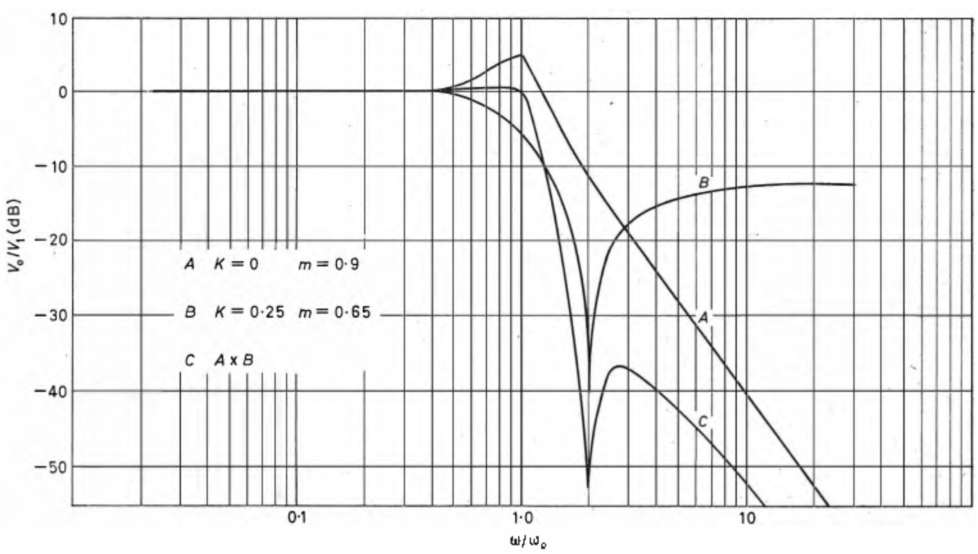


Fig. 8. Composite filter

variations of m at $K = 0$, and Fig. 7 for variations of K with m adjusted to give maximally flat response in the pass-band.

Composite Frequency Responses

By connecting several units of the basic arrangement, shown in Fig. 1, in series or parallel it is possible to achieve an extremely wide range of low-pass, high-pass, band-pass or band-stop filters. The values of m and K may be pre-set using fixed resistor networks to give the required pole and zero positions or may be made variable so that the final filter characteristic may be optimized for practical conditions. As an example of a composite characteristic two low-pass units A and B were connected in series and m and K of each unit set to the required value to meet the composite specification. Fig. 8 shows the characteristics for both A and B separate and the overall characteristic of A and B in series.

Conclusions

The analysis and experimental results agree closely and demonstrate how the simple circuit arrangement utilizing the familiar parallel-T network produces a pair of conjugate poles and a pair of conjugate zeroes. The loci of the poles on the p-plane is the arc of a circle of radius ω_0 centred about the origin, while that of the zeroes is along the frequency axis only. The locations of the poles and zeroes along their respective loci are controlled by two independent, non-interacting controls. The design and setting-up procedure of one of these basic units would be to design a symmetrical parallel-T network such that $\omega_0 = 1/RC$ where ω_0 is the desired filter cut-off frequency (-3dB point). Arrange the input impedance of the emitter-follower to be as high as possible and the output impedance of the feedback (mV_o), to the shunt-leg of the parallel-T, to be as low as possible. Similarly the output impedance of the potentiometer feeding KV_1 should be as low as possible. Set the control K to the required value as indicated by equation (6) so that maximum rejection is obtained at the desired frequency in the stop-band. Adjusting the input frequency to the cut-off value ω_o, trim control m such that the output is -3dB below the mid pass-band value.

Two Terminal Drive Circuits for a Piezo-Electrically Driven Mechanical Vibrator

By T. J. Stevenson*

The design of a circuit to drive a mechanical vibrator with a single piezo-electric transducer at resonance is complicated by the relatively large capacitance of the transducer. A suitable design is indicated by a consideration of the current in the transducer for a square wave voltage drive, and circuits are described which operate by changing the applied voltage between two levels when the current reverses sign.

As inductive elements would tend to resonate with the capacitance of the transducer, other methods are necessary if a drive of more than the supply voltage is needed. A suitable voltage trippler is described.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

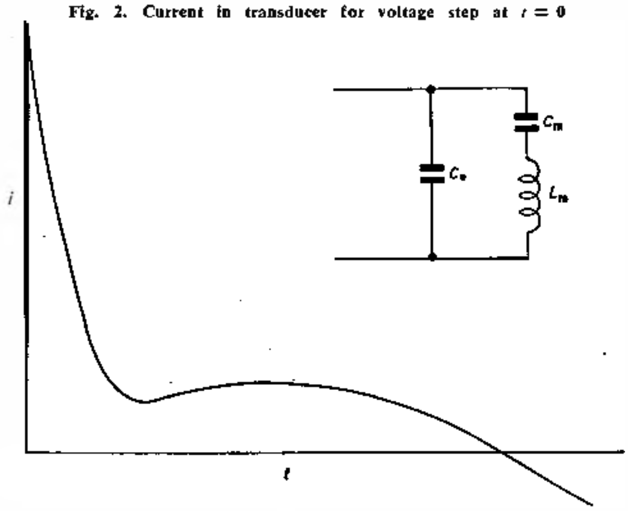
A CIRCUIT has recently been required to operate a vibrating unit (optical chopper) which is driven by a single piezo-electric transducer. The use of the more common system with two transducers, drive and pick-up, coupled with a fairly straightforward amplifier is precluded by mechanical considerations.

Since the mechanical system has a high Q, and its resonant frequency may be altered by effects such as temperature, it is vital that the frequency of the drive be controlled by the mechanical resonance alone. Now the equivalent circuit of the vibrator is shown in Fig. 1. C_e is the purely electrical capacitance of the transducer, while C_m and L_m represent the mechanical resonant circuit. (This is a somewhat crude analysis, but will suffice for the present purpose.) The problem was to devise a circuit which 'saw' resonance in C_m L_m without being swamped by effects in C_e. Another factor to be considered was that a peak-to-peak drive voltage in excess of the available supply voltage is needed. Any inductor or transformer will with C_e constitute a resonant circuit which may override the mechanical resonance and itself determine the frequency at which the circuit operates.

An approach which has been found useful in designing suitable circuits is as follows. Consider a voltage edge applied to the transducer terminals. The resulting current is shown in Fig. 2. There is a sharp pulse as the capacitance C_e is charged up (a small but not negligible source

Fig. 1 (inset). Simplified equivalent circuit of mechanical vibrator.

Fig. 2. Current in transducer for voltage step at $t = 0$



* University of Oxford.

impedance is assumed) and then a sinusoidal current due to the pick-up of the mechanical oscillations. The null in this current occurs at the extreme mechanical excursion—exactly when a reversal in the drive is required. So a circuit which applies a switched potential, and is controlled by a null in the current, will drive the vibrator at its resonant frequency.

The first circuit to be built is shown in Fig. 3. It is basically a simple bistable circuit (VT_1 and VT_4) with the transducer connected across the collectors. This push-pull arrangement gives an effective doubling of the supply voltage. The current in the transducer flows through either MR_1 and VT_3 or MR_2 and VT_2 according to its direction. But a current through VT_2 causes a current to the base of VT_1, which will tend to turn VT_1 on and VT_4 off. Now consider the case when VT_1 has just turned off, VT_4 turned on. The initial current will flow in MR_1 and VT_3, which will only tend to turn VT_4 harder on. But when the current reverses and flows in VT_2 and MR_2, VT_1 will be switched on, and the drive voltage reversed—which is what is required.

This circuit was found to have other modes of oscillation beside the mechanically controlled one described, but once started in the correct mode functioned very satisfactorily. It was not developed further because of mechanical changes to the vibrator which necessitated a higher sensitivity in the zero current sensing part of the circuit and a slightly higher drive voltage. To meet these requirements the circuit shown in Fig. 4 was designed.

The basic principle of the final circuit is shown in Fig. 5. This could be regarded as the simplest of negative resistance oscillators, but the circuit gain being far in excess of unity, it operates in a non-linear mode, and the applied drive is a good approximation to a square wave. Alternatively it can be understood with reference to Fig. 2. When VT_2 cuts off a current as in Fig. 2 flows in the base of VT_1, driving it on until the current falls to zero. When VT_2 then cuts on, a similar current of reverse sign flows through the path provided—MR_1 until is reverses and turns VT_1 on again. If R_2 is replaced by a large enough capacitor and clamping diode the system looks very much like an astable multivibrator. But its oscillations are still controlled by the mechanical resonance.

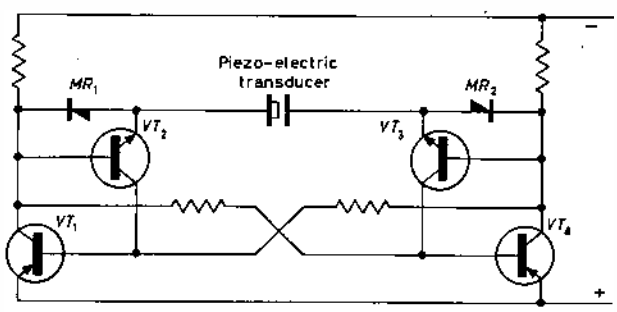
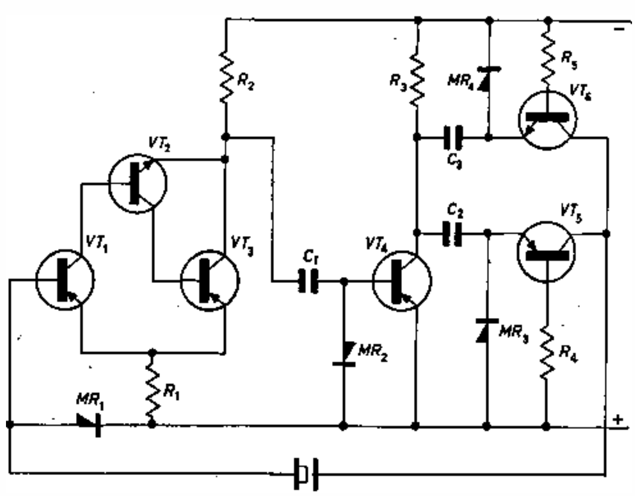


Fig. 3. First experimental drive circuit

Fig. 4. Final drive circuit



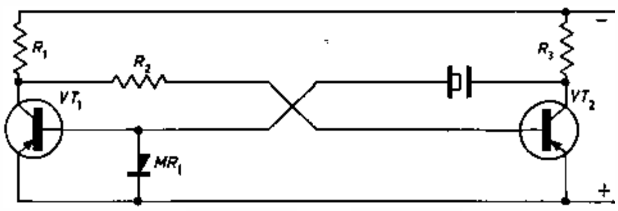


Fig. 5. Simplified version of final drive circuit

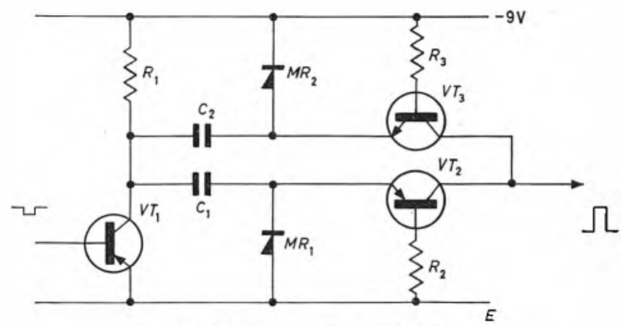


Fig. 6. Voltage trippling circuit

The circuit of Fig. 5 only gives a peak-to-peak drive voltage of the same value as the supply. To increase this the voltage trippling circuit shown in Fig. 6 was added. As the collector of VT_1 is switched between earth and the negative supply, the junction of C_2 and MR_1 is pumped 9V positive when VT_1 is on, and the junction of C_2 and MR_2 is pumped to -18V (-9V above the negative supply) when VT_1 is off. The output is switched between these values by VT_2 and VT_3. For example, when VT_1 is on, the emitter of VT_2 is positive with respect to its base resistor, a current flows in R_2, and VT_2 is on. But the emitter and the base of VT_3 are both at -9V and so it is off.

The transistors VT_1, VT_2, and VT_3 in Fig. 4 act as one transistor (or operational amplifier) of very high gain. Obviously low leakage silicon transistors must be used here, but otherwise components are not critical. Type numbers and component values have not been given as the design is highly variable to suit requirements, and the crudest rule-of-thumb calculations have proved adequate throughout.

Finally the history and *raison d'etre* of R_1 are amusing and instructive. It was included at an earlier stage on a purely bogus reasoning that an emitter-follower operation was desirable. When this was discovered and the resistor removed the circuit ceased to function. Without it the circuit responds to the falling edge of the first surge of current as C_e is charged (by some overshoot effect?) (Fig. 2.) The effect of R_1 is to nullify any remaining gain due to VT_1, VT_2 and VT_3 being driven harder and harder on in the bottomed condition. This is effected by the input current to the base of VT_1 flowing through R_1 and driving the whole transistor trio negative. This obliterates or reverses the gain for very high input, which prevents the edge mentioned having any effect.

Planning a V.H.F., F.M. Radio System for Bus Services with Computer Checking of Time-Tables

By T. K. Tawfig*, M.I.E.E.E.

In the system described a two-way communication system between a base station and buses is employed to provide information on the exact position of each bus in the service. The time and location of each bus is compared by computer with a stored time-table and the current position is displayed on a map in the control centre.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

AN ordinary two-way radio communication system used in a bus service for communication between the bus drivers and the central control office has no significant use alone, as such a system would require a large number of operators (instructors) at the base station. Also, such a system is rather slow in finding a solution to the difficulties introduced in a specified service due to deviation (delay) from the time-table. In such a case the operator has to study his own capability before proceeding with any plan to restore the normal running time. Further, in such a system the driver has to report his position in speech and no doubt, it takes time and effort to do so, plus the time required to establish calls by both sides. In view of the above disadvantages a more advanced and efficient system is required. A suggested system utilizing a computer is described below.

General

The system described is designed for use in a large city bus service. The system utilizes radiotelephone equipment which, although used for speech communication, is mainly used for checking continuously and automatically the positions of the buses running in the system and to read-out their assigned codes (numbers) and to compare the received information with a 'flowing' time-table which is planned for the service. Furthermore, the system has the facility of showing on a map the position of a bus which is out of the schedule running time by more than the permissible delay, also at the same time it displays its code on a digital display panel. Thus the system gives visual indication to the instructor about the trouble spots and the driver can then be contacted by radio so that the operator can introduce his solutions (Fig. 1).

System Description

The city or the town in which the service is to be installed is divided into parts or sections. Each section has a group of 'bus routes' running in it, say section 1, has the routes 1, 2, 3, 4 and so on. Now each route has its bus stops, the bus stops are numbered as bus stop 1, bus stop 2 and so on. This number is indicated on each stop and is illuminated by a lamp in such a way that the bus driver is able to see it, day or night.

This final bus coding will be made as shown in Table 1. Each bus is indicated by a 5-digit number. The first digit from the left indicates the section number. The next two digits indicate the route number on which the bus is running, the last two digits indicate the number of the bus stop at which the bus is at a given time, in a specified route. The first three digits are fixed while the last two are variable and changing while the bus is moving in the

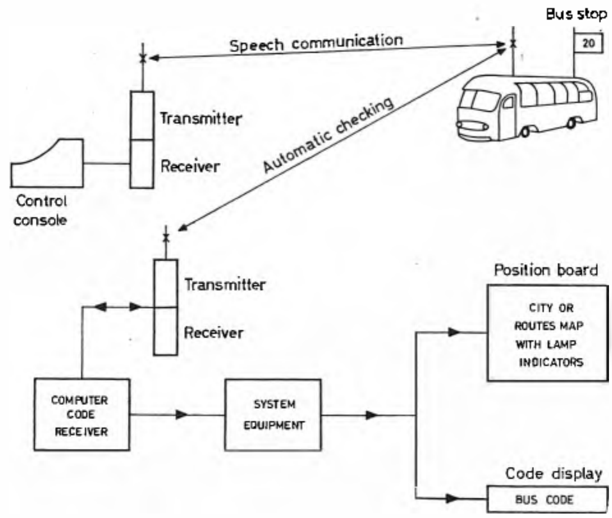
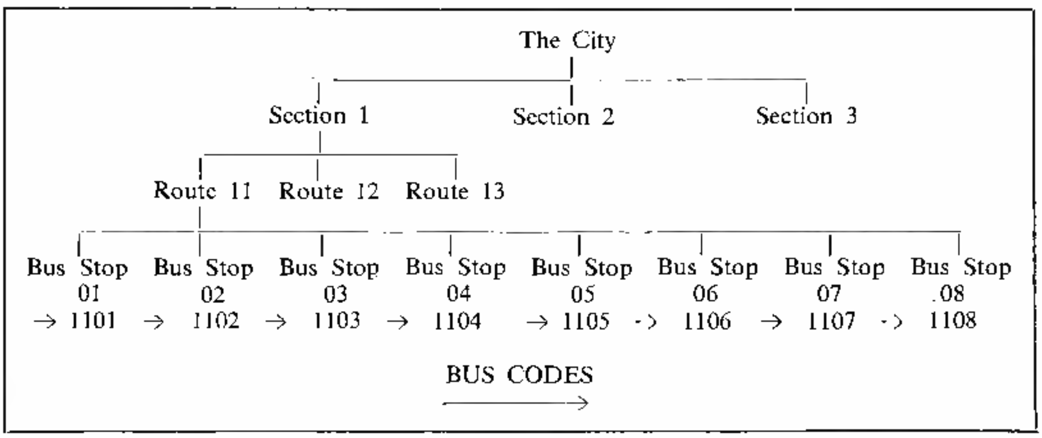


Fig. 1. The system general lay-out

network. For example, suppose at a given time the bus code was 5.88.22. This means that the bus belongs to section 5 of the city and it is running on route 88 and that at that given time it was at bus stop 22.

Each section of the city or town utilizes two duplex channels; one is used for speech communication and the other for the computer checking system. Each bus has a

TABLE 1



* *Storno Radiocommunication, Denmark.*

duplex mobile station which is controlled from a control box or panel which is placed under the dashboard and in front of the driver. The control panel has a front plate shown in Fig. 5.

The base station as shown in Fig. 2 has the following:

(1) A group of the radio-telephone stations with their normal control equipment. These are to be used for speech communication.
(2) A group of radio stations to be used with the computer system.
(3) A computer.
(4) Decoders, binary equipment and other auxiliary equipment.
(5) A digital-display unit, these are 5 or more digit display units. They are placed in front of the operators.
(6) One big map of the city or town or several maps showing parts of the city, i.e. showing a section with the routes shown clearly, also it shows the traffic lights existing along the routes. Small lamps are placed at the bus stops. These could be coloured lamps with a different colour for each route to ease identification.

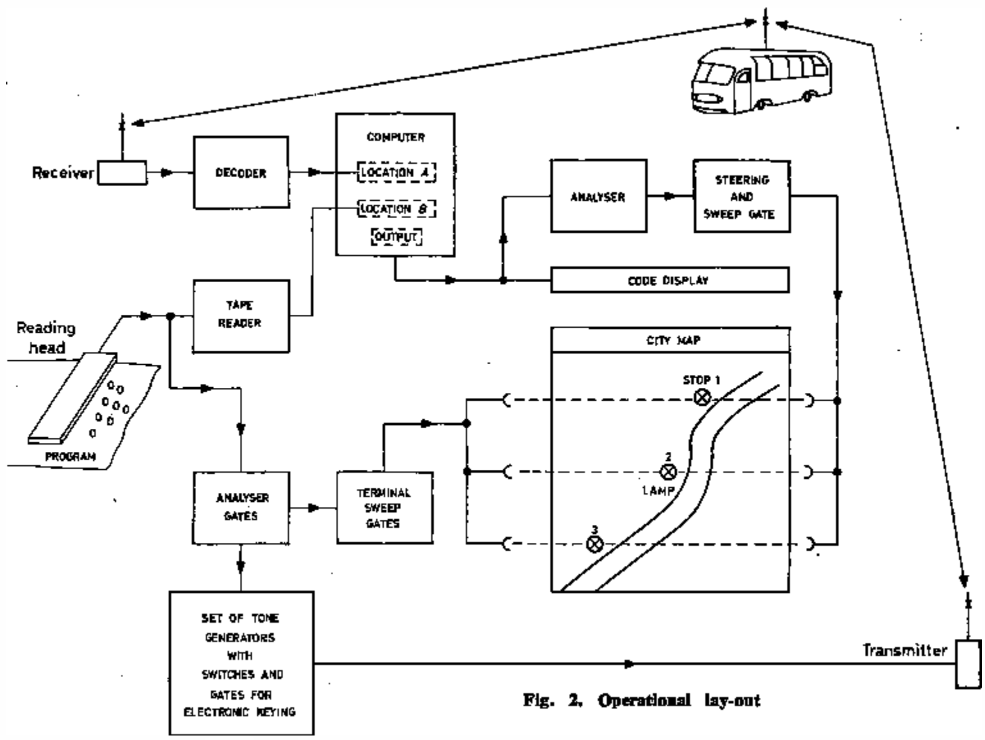


Fig. 2. Operational lay-out

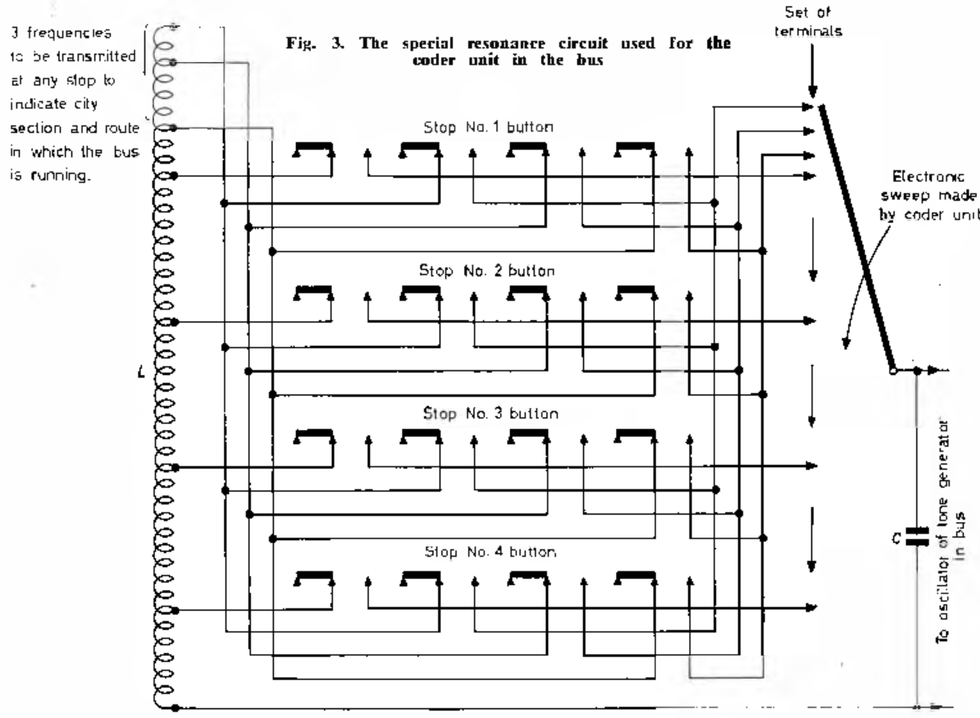


Fig. 3. The special resonance circuit used for the coder unit in the bus

Equipment Description

The Mobile Station in the Bus

This is a duplex station with channel shift facilities so that it has all the channels used for speech communication in the entire system so that the bus can be moved to any route. The mobile station, as was mentioned before, is controlled from the control panel. The latter has a special tone oscillator with a frequency determining circuit as shown in Fig. 3. Also it has a special tone receiver with switching circuits and gates (Fig. 4). The panel (Fig. 5) has rows of push-buttons marked 1, 2, 3 etc. On the front panel there is also a key switch which can be turned to one of three positions. The middle position will switch on the bus communication system to be used for automatic computer checking. The off position switches the entire system off. The speak position will allow the driver to use his station for speech communication. Furthermore, the front panel has four indicator lamps, these are:

(1) Key lamp, red.
(2) Call received, green.
(3) Tone on, yellow.
(4) Tone check, blue.

The key lamp lights when the transmitter is keyed.

The call-received lamp lights when the driver receives a call. The tone on lamp lights when the driver presses the bus stop buttons. The tone check lamp lights every time the computer checks the position and the code of the bus. The latter indicates to the driver that he can start moving from any stop so that he will be put in a mode of regular running which can be considered by the operator.

THE BASE STATION

This consists of the equipment shown in Fig. 2 and some of the main features of this equipment are as follows:

THE COMPUTER

The computer must have the following features:

(1) Sufficient number of characters.
(2) Sufficient locations in the storage.
(3) Punched tape programmed or film programmed.
(4) Output print facilities.
(5) Provide digital output suitable for displaying on a digital display unit.
(6) Time controlled programme running.

Other Binary Equipments

These employ switching and other logic circuits for automatic steering and logical analysing.

The Map

This is a large vertical board with routes clearly shown with the traffic lights and each bus stop is indicated by a lamp as described before, and one colour can be used for each route. The operator can switch off the automatically illuminated bus stop lamps after noting their position (Fig. 6).

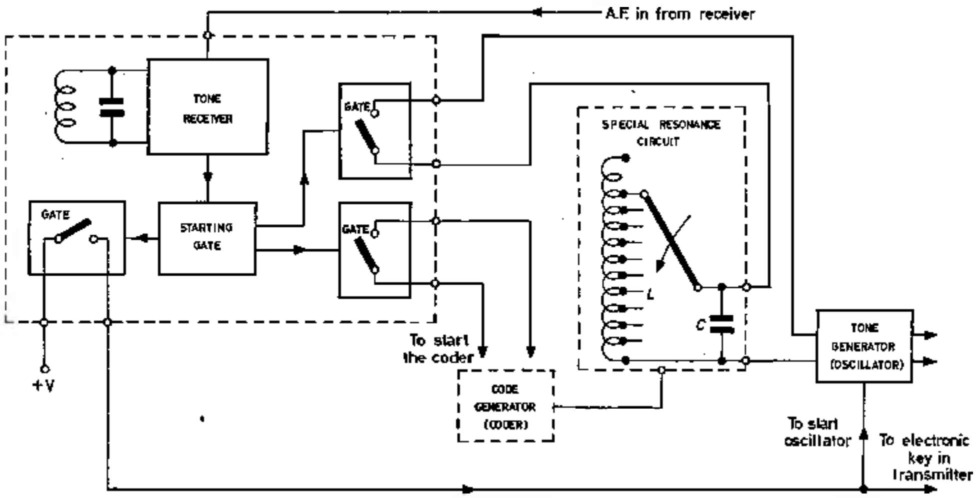


Fig. 4. Bus tone equipment system

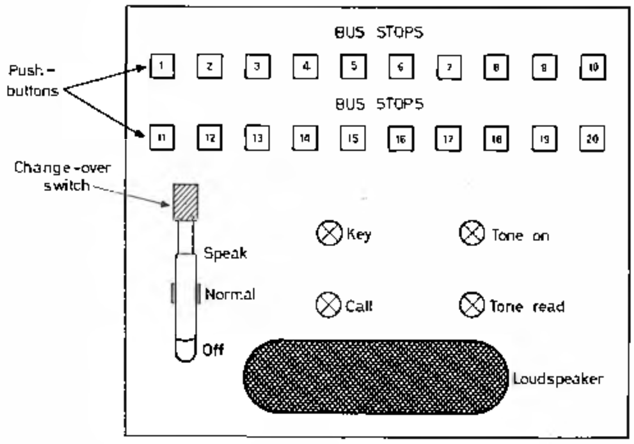
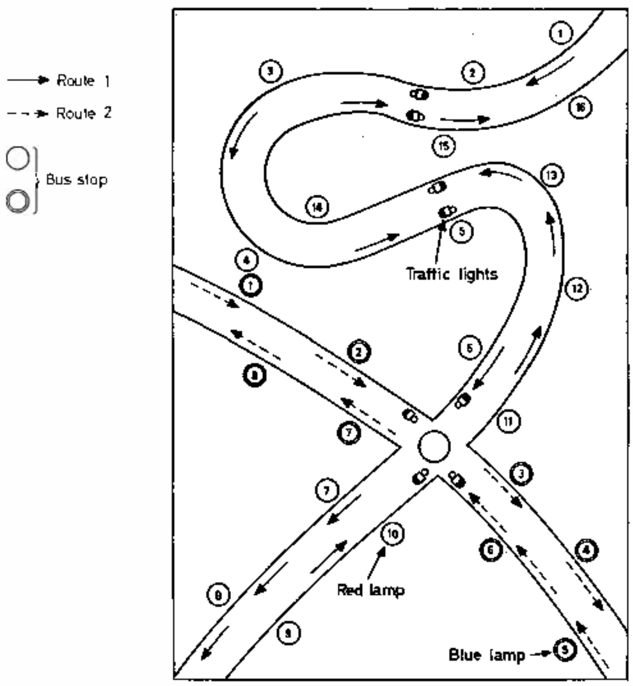


Fig. 5. The front plate of the control panel placed under the dash-board in the bus

Fig. 6. Example for the city or the routes map



System Mode of Operation

Every time the bus driver passes a stop he presses a push-button in his control panel which carries a similar number to that on the stop. This will connect a set of positions or taps on the frequency determining circuit of the tone oscillator to a set of terminals in the code generating unit. The bus receives a specified tone call at the specified time from the base station. This call is produced and transmitted from the base station as described in the computer system operation. However, this tone activates the tone receiver in the bus and its switching circuits. The latter will key the bus transmitters, switches and prepare the oscillator in the code generating unit, connect the frequency determining selected positions to the oscillator and start the code generating unit.

The latter unit will sequentially connect the prementioned selected taps or positions on the frequency determining circuit of the tone oscillator to the oscillator itself and therefore the oscillator will produce a set of sequential tones. The tones produced are fed instantaneously to the transmitter tone modulation inputs and will be transmitted from the bus to the base station. Now at the base station the received sequential tone combination is fed to the decoder which changes it to a sequential binary notation. This can be done by means of selective amplifiers, gates and diode matrix. The received information in the binary notation is then fed to the computer and is stored in one of its storage locations. The computer compares this information with stored information which is obtained from a running programme. The computer makes the comparison as shown by the flow chart in Fig. 7. If the bus under check was at the right place at the right time nothing will happen. But if the computer finds out by means of the programme that the bus was out of the planned or permissible running time limits, it would print out the received digital code of the bus and the time (this is for a study record), also it will feed the originally received digital information from the decoder to the output and to the digital display unit which will show the code of the bus concerned.

The binary output is fed to a binary system, described later, which switches on the lamp of the stop corresponding to the last two digits of the received code. Thus the operator will be able to see the code and the position of the bus. After notifying the stop at which the delayed bus is the operator switches off the illuminated lamp. The operator then uses the normal control box to call the driver. When the driver receives the speech communication call the green lamp will light so he switches over his system to the speak position and reports the condition and cause of the delay to the operator.

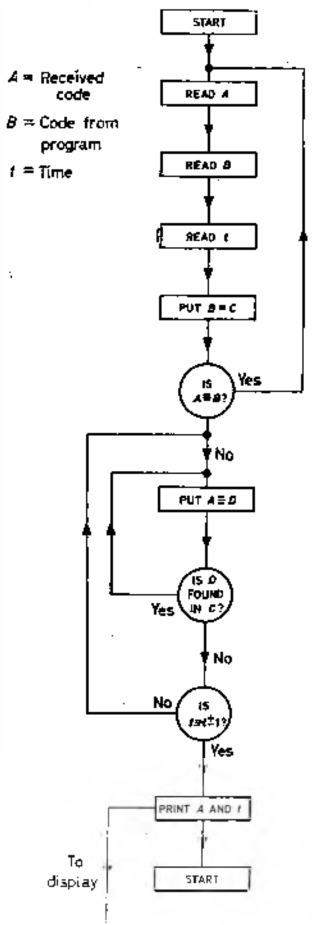


Fig. 7. Flow chart showing the operational mode for the computer

If the bus was late or delayed all along its route this will cause a continuous indication on the map and on the display of this bus due to continuous time table checking. To prevent this the operator asks the driver of that bus to press, starting from the next stop, the buttons with numbers higher by 1 or 2 than the numbers indicated on the next stop. This will put the bus in a delayed running time of one bus stage or two bus stages, but will preserve the accuracy of the checking system. If a bus is greatly delayed and to keep the services functioning properly, the operator can send an emergency bus; this can be any one of the buses used in the net, but with the key switch in the speak position, and such a bus should start service from a point or stop ahead of the trouble spot, by this means the running time is restored to normal all along the route.

Computer and Binary System Operation

Computer Programming

The procedures used in any programming are:

(1) Drawing the flow chart.
(2) Writing the programme using the computer code.
(3) Typing the programme on a teleprinter or any other device.

Computer and Binary Operation

As shown in Fig. 7 the computer receives the code from the decoder in binary form and will store this in location *A*. Also it will receive at the same time a delayed code from the programme and will store this in location *B*. The delay being to compensate for the time taken by the checking method. The computer then reads the time from the programme and stores the information in location *t*. The computer puts all the received codes from the programme in another location, location *C*. The computer then asks itself are the codes *A* and *B*, which are received at the same time, equal? If they were, the computer proceeds in receiving the next code, if not, the computer will put the first received *A* to location *D* and now it asks itself, if *D* can be found in *C*? If the answer is yes,

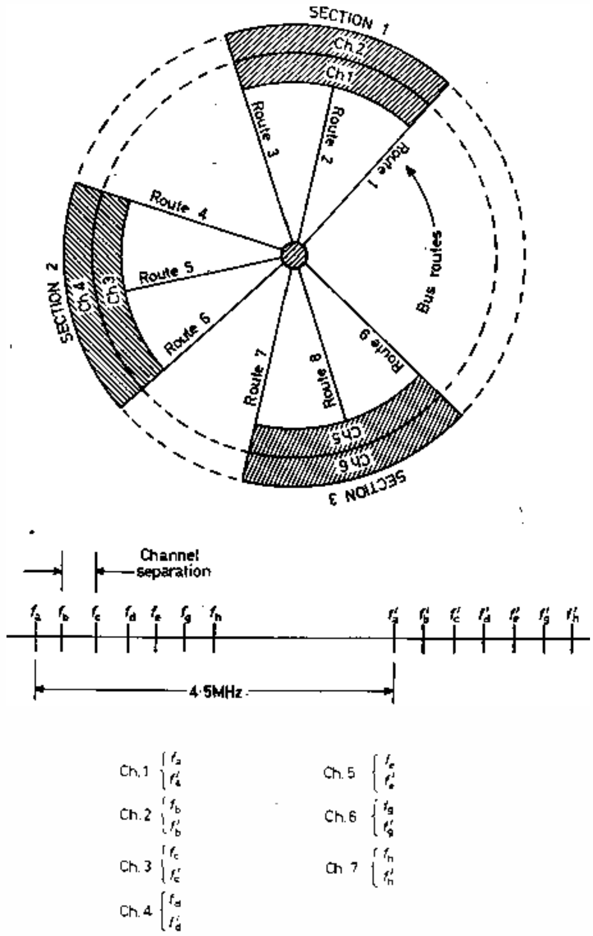
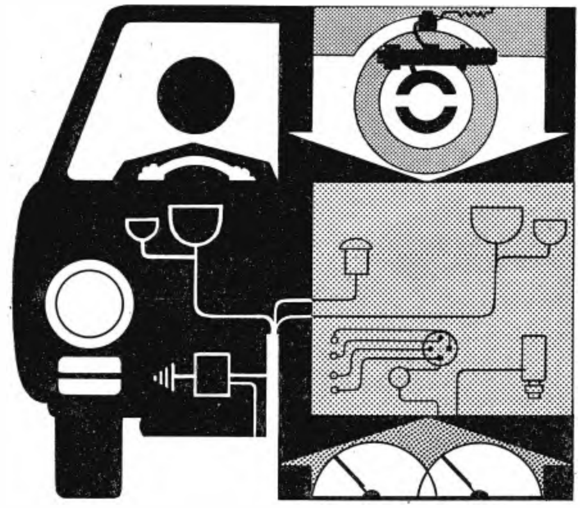


Fig. 8. Sketch showing how the available frequency channels can be distributed

The frequency scale shows the frequency lay-out

the computer proceeds as before, if no, the computer asks whether the time is equal to *t*, say 1 minute, if the answer is no, the computer proceeds as before, if yes the computer will print *A* and the time, also it will feed this *A*, which is in binary notation, to a display unit which can show the code. Also this binary code is fed to a gate with frequency dividers and gates which can analyze the first digits, i.e. the 'section' digit and will analyze the next two digits and direct the remainder to the gate and the matrix of the 'stops'. This will direct a pulse to the corresponding 'stop' lamp which will light and stay so until the operator switches it off.

Now at the same time the information from the programme is read and fed to the computer; the code is fed also to a gate and to a steering matrix similar to the one described above. This will then activate the particular or corresponding tone generator instead of lighting a lamp to produce the tone call which is then transmitted to the bus for checking its conditions as was described before.



ELECTRONIC EQUIPMENT FOR MOTOR VEHICLES

By G. N. Patchett*, B.Sc., Ph.D., C.Eng., F.I.E.E., M.I.E.R.E., M.I.E.E.E.,

Electronic devices have not been applied to motor vehicles to any extent. Valves are difficult as the high anode voltage required is not readily available but semiconductors fit in well with vehicle supplies. Several devices are described in this article which are completely solid state and based on making the vehicle safer. Three are designed to ensure that the lights are operating as required by law; both brake lights, all four side lights and side lights at lighting up time. One helps to ensure that the brakes are in order and another to reduce the possibility of accidentally exceeding the speed limit. The final device is designed to improve vision under intermittent rain conditions.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

THE design of electronic equipment for cars and other motor vehicles is not easy as the supply voltage varies considerably and the temperature variations are large compared with those encountered in many other applications. The voltage is lowest when the engine is stationary and either a large load such as headlamps or heater is switched on or a smaller load such as side lights has been on for a considerable time. The maximum voltage occurs when the battery is being charged and is almost fully charged. The maximum voltage depends on the condition of the battery and the setting of the automatic voltage regulator. For the purposes of the design the voltage range has been taken as 10 to 16V which is rather larger than would normally be expected. Regarding the temperature range the minimum temperature has been assumed to be 0°C although it is realized that if the car has been stood outside for some time it may be lower than this. The maximum temperature depends on the siting of the device which obviously should be as far away as possible from the hot engine and particularly the exhaust system. As well as the heat from the engine the temperature of any device can be quite high if the car is exposed to the sun for some time. The maximum temperature has been assumed to be 45°C i.e. a temperature range 0° to 45°C.

Brake Light Indicator

It is very desirable that both brake lights of a vehicle are operating correctly. It is surprising what a delay occurs before a driver realizes that a vehicle in front is slowing down when the brake lights are not operating and a single brake light is easily confused with a trafficator. The device to be described lights an indicator lamp only if *both* brake lights are operating. An electromagnetic relay may be used for this purpose but the design is rather difficult due to the initial surge of current in the lamp circuit. Such a device was used by the author some years ago and is now a commercial product. It was considered that an electronic device would be more satisfactory as relay contacts tend to be unreliable unless a sealed relay is used.

The basic principle of the electronic device is shown in Fig. 1. In all the circuits described a positive earth supply is assumed but the circuits can be changed to negative earth by suitable change of transistors. A low value resistor R_1 is placed in the brake light circuit so that when the brakes are operated (and the brake light switch is closed) a suitable voltage is applied between base and emitter of the transistor VT_1. This causes a current to flow through the indicator lamp L_1. If one of the brake lights fails the voltage across R_1 is not high enough to cause the flow of base-emitter current and hence no collector current flows in L_1.

The circuit is unsatisfactory as it stands because the transistor would be destroyed on first operating the brakes, due to the large surge of current through the brake lamps when the filaments are cold. The brake lamps are normally 21W each and hence the steady current on 12V is 3·5A. The cold resistance is only approximately 0·32Ω each or 0·16Ω for the two in parallel. With no resistance in circuit the initial current would be $12/0{\cdot}16 = 75$A! The brake light circuit has some resistance and this will vary from vehicle to vehicle and depend on the condition of the brake light switch contacts, lamp contacts etc. but appears to be of the order of 0·3Ω. Assuming this resistance and neglecting R_1 of Fig. 1 the current is still $12/0{\cdot}46 = 27$A. The transistor must therefore be protected from this current surge.

The base-emitter voltage of the transistor VT_1 (silicon) is about 0·7 to 0·75V, for full current through the indicator lamp L_1. By making $R_1 = 0{\cdot}24\Omega$ the normal voltage drop with two lights is $0{\cdot}24 \times 3{\cdot}5 = 0{\cdot}84$V which will operate

* *University of Bradford.*

the transistor. The worst surge condition is when the supply voltage is 16V and under these conditions the voltage across R_1, assuming a cold lamp resistance and wiring resistance of 0·46Ω, is $0{\cdot}24/(0{\cdot}24 + 0{\cdot}46) \times 16 = 5{\cdot}5$V. This would obviously destroy the transistor. The transistor is protected first by placing a diode *MR* across R_1 as in Fig. 2. The diode will now limit the rise of voltage across R_1 but must be capable of passing the large surge current without damage. A Ferranti ZS70 has been used which gives a voltage drop of about 2V with a current of 30A. Allowing for this 2V drop the initial current with a 16V supply is $(16 - 2)/0{\cdot}46 = 30$A.

It is still necessary to limit the base current by resistor R_2. Assuming a base-emitter voltage of 0·85V under the conditions of maximum base current (100mA for a BFY52) the resistor $R_2 = (2 - 0{\cdot}85)/0{\cdot}1 = 12\Omega$, a value of 20Ω is used to be on the safe side.

It is now necessary to check that the device will operate over the supply voltage range and will not operate with one lamp only. The current of the lamps is not proportional to voltage but varies from 3·16A to 4A for a voltage variation of 10 to 16V. The drop across R_1 under steady conditions (neglecting the current in *MR*) is therefore 0·76 to 0·96V and sufficient to operate the transistor. As only about 2mA base current is required to give full current in L_1 the drop in R_2 can be neglected being only 40mV. With only one brake lamp operating the current at 16V (the worst condition) is 2A and the drop across $R_1 = 0{\cdot}48$V. This is far too low to cause appreciable base-emitter current and hence no collector current flows in L_1. The initial surge of current with one lamp is sufficient to give a short duration flash of the indicator lamp but this can be prevented, if desired, by a capacitor C_1 between base and emitter of VT_1 as shown dotted in Fig. 2. The capacitor should be at least 500μF but 6V working is sufficient.

Variation of temperature will, of course, cause variations of base-emitter voltage drop for a given base current and this is of the order of ±0·05V from that at 20°C for the range 0°C to 45°C. Even allowing for this it will be seen that the circuit will still operate satisfactorily over the full voltage range.

The indicator lamp L_1 should be as low consumption as possible subject to sufficient light and a 2·2W is used. Although the power dissipated in the transistor VT_1 is zero with zero current and small with full current (when the voltage across it is low) the transistor must change from one condition to the other and it is difficult to be certain that under *all* conditions the transistor will not operate approximately half on. Under these conditions the dissipation is a maximum. The dissipation is a maximum with a 16V supply and with 10V across the transistor (this being obtained from the lamp characteristics, the maximum dissipation not taking place at half voltage due to the non-linear characteristics of the lamp). Under these conditions the dissipation is 900mW and in order to operate at 45°C a BFY52 requires a heat sink. The indicator lamp has, of course, a low resistance when cold and there is a current surge on switching on. The cold resistance is about 10Ω and hence the current with a 16V supply is 1·6A which is excessive for this transistor (1A being the maximum). This surge is prevented by a 6Ω resistor R_3 in series with L_1 which causes a negligible drop under normal conditions.

An alternative circuit is shown in Fig. 3, this having the advantage of reducing the surge current in the diode *MR*. The drop across *MR* is now about 1·3V and hence a lower value of base resistor (R_4) may be used. It is necessary to increase the value of R_1 slightly to allow for the drop in R_2 due to the diode current.

A device of this type should be designed to fail to safety and as far as possible this is the case. Failure of bulb L_1 or an open circuit in VT_1 or R_2 or R_3 of Fig. 2 will cause the bulb not to light and indicate that something

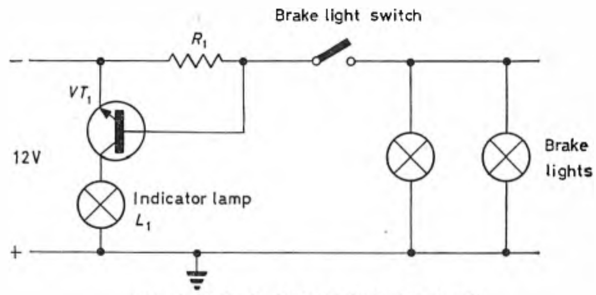


Fig. 1. Basic brake light indicator circuit

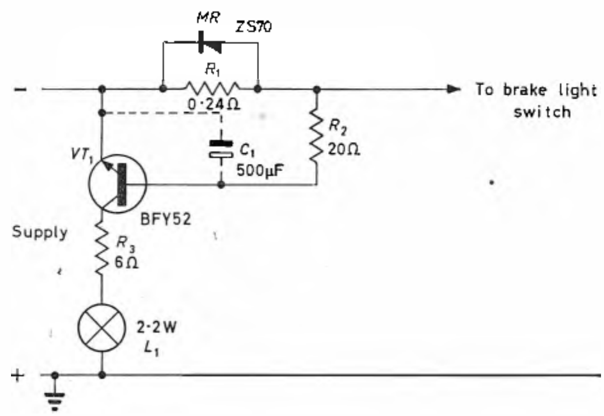


Fig. 2. Practical brake light indicator circuit

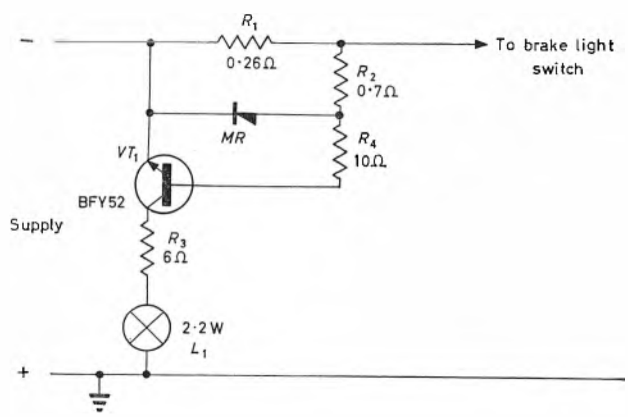


Fig. 3. Alternative brake light indicator circuit

is wrong. A short-circuit in the transistor will cause the indicator lamp to glow continuously indicating a fault. It should be noted that the device is placed before the brake switch for this reason.

Side Lamp Indicator

The purpose of this device is to indicate that all four side lights are operating.

One sees far too many cars with one side light off, a condition which can be extremely dangerous particularly if the off-side light.

The number plate light will be neglected for the present. It sounds relatively easy to check that all four lights are operating by measuring the total current but this is complicated by the variations in supply voltage. The side lights are normally 6W each. The total current on 12V is therefore approximately 2A. The variation of current for a voltage range 10 to 16V is 1·6 to 1·96 (using particular lamps which all were below 6W although marked 6W). The current of three lamps on 16V is 1·47A and this is too close to 1·6A when allowance is made for bulb tolerance etc.

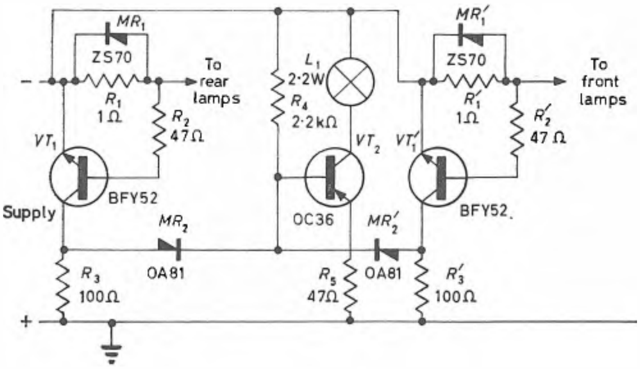


Fig. 4. Side light indicator circuit

The principle used is the same as that for the brake lights, the side lights being divided into two circuits—the front two and the rear two. The current of each pair is checked by a circuit similar to the brake light circuit and the outputs from these circuits are then fed to an AND gate. This will only operate when both inputs are present. The circuit is given in Fig. 4. The value of R_1 and R_1' is increased to allow for the smaller current of about 1A instead of 3·5A. The initial current surge is less, being about 13A, allowing for a circuit resistance of 0·3Ω. R_2 and R_2' can be increased in value as the transistor collector current is now smaller. If both pairs of lamps are operating then VT_1 and VT_1' are conducting and a voltage of some 11V (on a 12V supply) is developed across R_3 and R_3'. Diodes MR_2 and MR_2' are reverse biased and the current through R_4 flows into the base of VT_2 causing collector current to flow in L_1, indicating all four lamps are operating. If any lamp fails then the current in VT_1 or VT_1' is cut off, diode MR_2 or MR_2' becomes conducting and shorts the base of VT_2 to earth through R_3 or R_3', thus cutting off the collector current. An OC35 or OC36 is used for VT_2 as the dissipation may reach 900mW. No heat sink is required. The dissipation in VT_1 and VT_1' is now less and no heat sinks are required.

The indicator lamp will go off should there be a failure of one or more bulbs or a failure anywhere in the side lamp circuits.

Unfortunately the rear number plate light is usually wired to the rear side lights and in the circuit given it is necessary to feed the number plate lamp separately. An alternative is to feed the two rear lamps and number plate lamp through R_1 (Fig. 4) but reduce the value of R_1 to about 0·7Ω to allow for the increased current. The circuit must now discriminate (under the worst conditions) between the current of three lamps at 10V which is 1·18A (using particular lamps) and two lamps (fault condition) at 16V when the current is 0·98A. This is rather marginal if allowance is made for lamp tolerance, transistor tolerance and temperature. By careful adjustment of R_1 to suit a particular transistor it is just possible to get the circuit to operate and this may be satisfactory particularly when it is considered that a rather large variation of supply voltage has been assumed.

Automatic Operation of Side Lights

This device switches on the side lights when the ambient lighting reaches a predetermined value, corresponding approximately to lighting up time. This is particularly useful when the car is parked during daylight and the owner is not certain how long he will be away.

Although a photocell and relay may be used it was decided to make the circuit completely electronic. The circuit is shown in Fig. 5. The voltage across the photoconductive cell C is compared with the voltage across a portion of R_5, so that the circuit is reasonably independent of supply voltage. Transistor VT_1 acts as an emitter-follower (R_2 only limiting the maximum current in VT_1) so that a high impedance is presented to the photocell. Transistor VT_1 feeds transistor amplifiers VT_2 and VT_3 which control the main series transistor VT_4, which is connected as an emitter-follower. With bright light the resistance of the cell C is low and hence the voltage across it is low and, if less than that across the lower portion of R_5, transistor VT_1 is non-conducting. Similarly VT_2 and VT_3 are non-conducting and the base of VT_4 is at earth potential and no voltage is applied to the side lights.

At a certain value of ambient light on the cell the voltage is sufficient to cause VT_1 and VT_2 to conduct. This causes VT_3 to conduct heavily and the base of VT_4 is approximately at negative line voltage. Since VT_4 is an emitter-follower the emitter follows the base voltage and normal voltage is applied to the side lights. It is undesirable for the side lights to come on slowly as this is likely to be confusing and also the dissipation in VT_4 is high (up to 9W). To prevent this the circuit is made to trigger by the use of the feedback resistor R_8. With bright light the right-hand side of R_8 is at earth potential (positive line). When the cell starts to conduct so does VT_3 and hence the right-hand side of R_8 moves in a negative direction so that a current flows in R_8 causing VT_1 to conduct more. Thus the circuit switches on rapidly. A similar action takes place on switching off.

The potential divider R_5 acts as a sensitivity control, the greater the voltage across the lower portion of R_5, the greater must be the voltage across the cell to switch

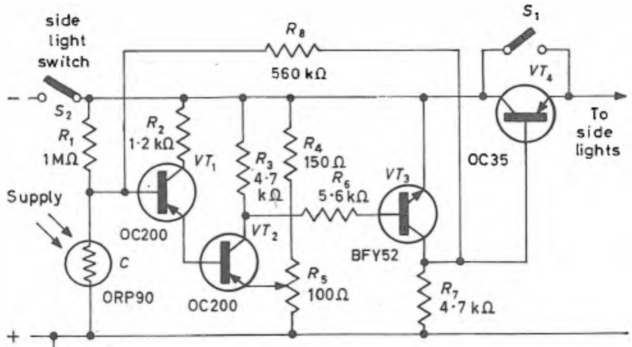


Fig. 5. Circuit for automatic operation of side lights on reduction of ambient lighting

the circuit on, this meaning that the ambient light is less. It is important to prevent an excessive surge of current in VT_4 on switching on and this is prevented by limiting the base current of VT_3 by R_6, which in turn limits the base current of VT_4. Assuming a maximum voltage across R_6 of 9V the base current of VT_3 is 1·6mA. If the current gain of VT_3 is 60 the maximum collector current and base current of VT_4 is 60 × 1·6 = 96mA. Assuming a current gain of VT_4 of 50 the current is limited to $96 \times 50 = 4800$mA $= 4{\cdot}8$A which is well within the current rating of VT_4. The normal dissipation in VT_4 is approximately $1 \times 2{\cdot}5 = 2{\cdot}5$W, assuming a drop of 1V and a lamp current of 2·5A. Thus VT_4 must be mounted on a suitable heat sink. In order to prevent false operation by car headlights and street lights the photocell should be mounted so that direct light from such sources will not fall on it. When the device is not being used VT_4 is shorted out by switch S_1 and put into

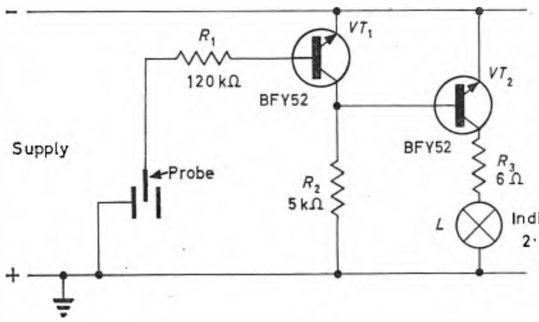


Fig. 6. Brake reservoir indicator circuit

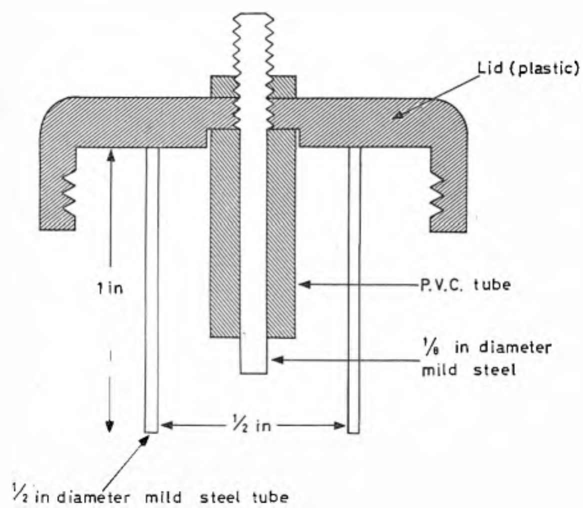


Fig. 7. Construction of brake reservoir probe

operation by opening switch S_1 and of course, closing the normal side light switch.

As the circuit is shown in Fig. 5 there is about a 3 to 1 difference between the light to switch on and that to switch off. This may be reduced by increasing the value of R_8 and when R_8 is 2·5MΩ the difference is about 30 per cent. It is important that R_8 is not increased to a value which will reduce the feedback so that a rapid switch over does not take place. There is some variation of the light value to switch on or off due to change of voltage but as the voltage will normally be approximately constant when the device operates this is not important.

There are some advantages in making the circuit so that once the cell has switched the lights on they can only be extinguished by switching off the circuit by the normal side light switch. This can be done by removing R_8 and connecting the lower end of the cell (through a limiting resistor of say 15kΩ) to the base of VT_4. Once the circuit operates the lights are maintained on by the current through the cell.

The same circuit may be used for a parking light but in this case VT_4 can be dispensed with and R_7 is replaced by the parking lamp (not exceeding 2·2W). Depending on the current gain of VT_3 it may be necessary to reduce R_6 to say 4·7kΩ.

Brake Reservoir Indicator

It is very desirable to have some warning device to indicate a drop in brake fluid level in the brake reservoir. A drop in level indicates a leak and if not attended to the brakes will eventually fail to operate. The device is required to light a warning lamp if the level falls below a predetermined value. A device of this nature has on two occasions given warning of a leak to the author and the leaks might have had fatal results if they had not been attended to.

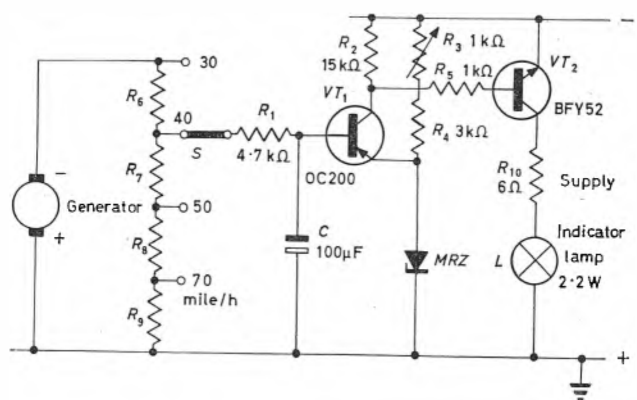
A float may be used with contacts which close when the level drops. This is not very reliable as only a small float can be used (due to the size of reservoir) and hence only light contacts can be used which tend to be unreliable, the device not failing to safety. The author has tried a number of ways of measuring the fluid level from heated thermistors to heated filaments but all these failed due to the large range of ambient temperature involved together with large variations of supply voltage.

The present device operates on the conductivity of the brake fluid which is quite high. The circuit is shown in Fig. 6. Provided the probe is in the fluid a current flows in R_1 and in the base of VT_1. This results in a low base-emitter voltage on VT_2 and hence VT_2 does not pass sufficient current to light the indicator lamp L. If the fluid level drops no current flows in VT_1 and VT_2 conducts due to the current in R_2 flowing into its base. As it is undesirable to pass a large current through the probe and the brake fluid, R_1 is made of high value and limits the current to about 100μA. The lamp L remains lit with a current of only about 20μA. The probe should be constructed so that the leakage path is large as otherwise the device will operate on the leakage current over the insulation which gets covered with brake fluid. A suitable construction is shown in Fig. 7 this fitting on to the screw cap of the reservoir. The probe is surrounded by a metal cylinder so that the current flows to it in preference to other parts of the braking system.

From the author's experience this small current does not appear to have any detrimental effect on the brake fluid and much larger currents have been used by the author without any known deterioration of the braking system. It is difficult to be quite certain that there is no effect over a long period under all conditions. The effect can certainly be made negligible by either:

(a) Arranging to energize the circuit through a push-button which is pressed when it is desired to check the level of the brake fluid. This is not as satisfactory as it depends on the driver making regular checks but is better than nothing.

Fig. 8. Circuit of speed limit indicator



(b) Connecting the circuit across the brake light so that it is only energized when the brakes are applied which is a small percentage of the total driving time. If the Brake Light Indicator is used then the two indicator lamps can be placed near together. If everything is correct then only the brake light indicator light should operate, operation of both lights (or no lights) indicates a failure which should be investigated.

It is difficult to make such a device so that it will fail to safety under all conditions but it does fail to safety if there is a break in the wiring anywhere between R_1 and the probe and due to an open-circuit is R_1 or VT_1. If the bulb fails it does not fail to safety and a check button may be added which opens the circuit between R_1 and the probe. An alternative is to alter the circuit so that the lamp is normally on but this is not considered desirable as a driver is more likely to notice a light coming on than one going off. In fact all other indicator lamps on a car operate in this way, e.g. oil pressure light, ignition light.

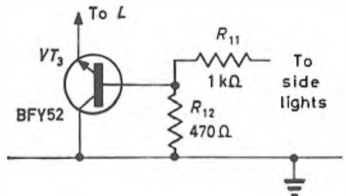


Fig. 9 (left). Circuit for reduction of brilliance of indicator light.

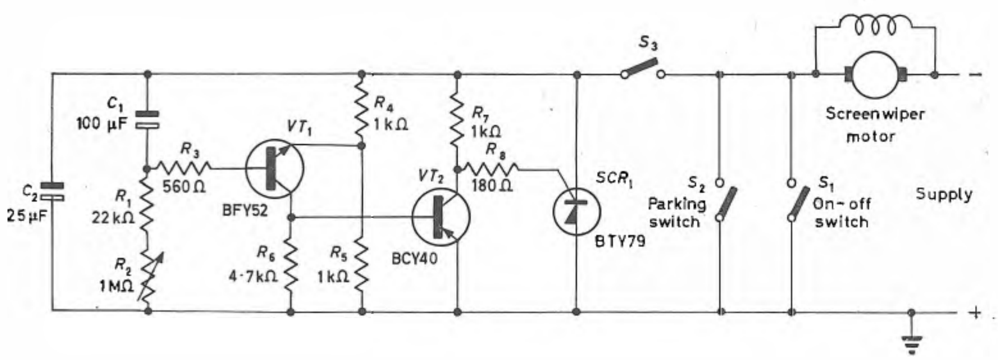


Fig. 10 (below). Screenwiper control unit

Speed Indicator

This device is designed to light a lamp when the car speed exceeds a predetermined value; the predetermined value being switched to the present four speed limits of 30, 40, 50 and 70 mile/h. The circuit is shown in Fig. 8. A small permanent magnet generator is fitted in the speedometer drive cable so that a voltage is obtained proportional to vehicle speed. The generator should preferably give a voltage of 3 to 6V at 30 mile/h. The one used by the author gave a voltage of about 3V. Transistor VT_1 is normally non-conducting as the emitter is biased negatively by the Zener diode *MRZ*. Hence VT_2 is non-conducting and the indicator lamp *L* is not lit. When the speeds reach a certain value VT_1 conducts, causing VT_2 to conduct and light lamp *L*. The potential divider R_6-R_9 together with switch *S* is used to select one of the four speed limits. R_1 limits the current in VT_1 and *C* is used to remove any commutator ripple from the generator voltage. The voltage of the Zener diode is chosen to fit the particular generator and fine adjustment of operating speed may be obtained by R_3. It is desirable to choose R_6 to R_9 so that the indicator lamp operates slightly above the speed limits say 33, 43, 53 and 73 mile/h so that *L* is normally out. As a simple amplifier is used rather than a trigger circuit the brilliance of the lamp changes gradually over about 5 per cent speed change. This was found to be more desirable in practice than a sudden change.

Variation of supply voltage and temperature do cause some change in the precise speed at which the device operates. The magnitudes of the change will depend on the Zener diode characteristics. During the times when the device is operating the supply voltage will not vary by large amounts (i.e. the engine will always be running and charging the battery) and it was considered accurate enough.

At night time it is desirable to reduce the brightness of lamp *L* and this may be done by an additional transistor as in Fig. 9. When the side lights are off transistor VT_3 has little effect but with the side lights on the base is made negative and hence there is a drop across VT_3 depending on the ratio of R_{11}/R_{12}.

Rather than have the indicator lamp on the instrument panel it is better to position it so that it is reflected on the windscreen. In this way it is almost impossible to drive without one's attention being drawn to it. As it should not normally be lit it does not interfere with vision through the windscreen.

Screenwiper Control Unit

There are times (during drizzle) when one does not want a screen wiper to operate continuously, but only intermittently and the circuit was devised for this purpose. The circuit is given in Fig. 10. Switch S_1 is the normal screenwiper on-off switch and S_2 the internal parking switch on the wiper motor which opens when the blades reach the parking position. S_3 is the on-off switch for intermittent operation which is conveniently combined with the timing control R_2 (which controls the times between wipes).

On closing S_3, the 12V supply is applied to the circuit through the screenwiper motor which does not run due to the small current. Transistor VT_1 is cut off since capacitor C_1 is initially discharged. Hence VT_2 is non-conducting and the gate of SCR_1 is returned through R_7 and R_8 to the cathode and hence SCR_1 is non-conducting. After a time interval depending on the value of R_2, C_1 has charged sufficiently to cause VT_1 and hence VT_2 to conduct. The gate of SCR_1 is now connected through R_8 (a limiting resistor) and VT_2 to negative line and hence SCR_1 is fired. This causes the screen wiper motor to run but soon after moving from the parked position S_2 closes, this switch shorts out the circuit and SCR_1 becomes non-conducting again. On returning to the parking position S_2 opens and the motor stops until C_1 recharges again. Due to the motor inductance it is necessary to place C_2 across the circuit to limit the rate of rise of voltage or SCR_1 fires each time S_2 opens. As the s.c.r. is carrying the full current of the motor and the starting current (about 10A) it must be mounted on a suitable heat sink. The circuit does not in any way interfere with the normal operation of the wipers by S_1, which over-rides the intermittent operation, if in use.

All the devices described have been used successfully for some time on a car without any failure.

Digital Coding Equipment

for Recording Field Strength Measurements

By D. G. Beadle*, B.Sc.(Eng.), A.C.G.I., C.Eng., M.I.E.E., and D. E. Susans*, C.Eng., M.I.E.E., A.M.I.E.R.E.

The article describes equipment which has been developed to translate the d.c. output of one or more field strength measuring receivers into digital form. The outputs of the receivers are sampled at regular intervals and the values recorded by means of a 5-unit binary code on to a paper tape which can subsequently be fed into a computer for analysis.

Results are given which show good agreement with other methods of analysis.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

WHEN planning a broadcasting service, it is essential to know the degree to which the aerial will be marred by interference from other stations. For a number of years investigations, co-ordinated by the International Radio Consultative Committee (CCIR), have been conducted on how the field strength for various percentage times varies with frequency and with distance from a transmitter. Propagation curves are derived from the data collected over long periods of time by organizations throughout the world. As a contribution to this work the BBC has set up a number of unattended receiving stations where the field strengths of various transmissions are recorded. For example, for the u.h.f. and v.h.f. bands, the sites have been chosen so that the distance from the transmitters falls within the range 60 to 250km. The data are normally collected in the form of roll charts from a pen recorder depicting field strength plotted against time. These charts then have to be analysed to determine the total time for which a given level of field strength has been exceeded, a process which is both laborious and time-consuming. Some more efficient method would clearly be a great advantage.

The receiver that has been used for this work gives a d.c. output corresponding to the logarithm of the field strength. Equipment, consisting of a voltage coder and a tape perforator, has been developed to sample the output voltage of one or more receivers at regular intervals and record the values in coded form on a paper tape. The punched tape is of the standard type so that it can subsequently be fed into a digital computer for analysis. The 0 to -3V output of a receiver is coded into thirty steps of 100mV each (thirty-one levels). The thirty-second character of the five-unit binary code, the 'all-blank', is used as an identification of the start of the sampling cycle of a group of receivers. Sampling rates of one, two or three times per minute may be obtained by the use of interchangeable cams. These rates were chosen so that there would be sufficient samples to give the required statistics of the signal with the same accuracy as those from a chart record.

It is advantageous to operate conventional chart recorders in parallel with the coder unit for the following reasons:

(a) They provide an alternative method of analysis should a fault develop on either the coder or perforator.
(b) A check may be made at any time on the accuracy of the coder results.
(c) Visual inspection provides information on fault conditions at the transmitter and reveals any unusual characteristics that may arise through interference.
(d) Certain information, e.g. fading rates, times of abnormal propagation, variations of hourly medians may be more readily obtained from recorder charts than from computer analysis of paper tapes. This, of course, depends on the sampling rate and on how comprehensive the computer analysis programme can be made.

A standard reel of paper tape is approximately 300m in length and this will record the output of a field-strength measuring receiver operating for fourteen hours a day for more than three months. It is convenient, however, to collect and analyse the tape once a month on a digital computer; this ensures that a fault does not go undetected for too long. At present, the analysis is confined to the determination of the number of samples recorded for each

Fig. 1. (above) Photograph of coder and tape perforator showing multi-channel input unit, coder with single channel input unit and tape perforator

* *British Broadcasting Corporation.*

level and hence the percentage time that a given level of field strength has been exceeded for a particular distance from the transmitter.

The computer takes only ten minutes to analyse a month's recording for all the levels, while by the old method of chart analysis it would take five to eight days depending on the depth and rapidity of fading. A considerable saving in man-hours is thus effected by the use of automatic analysis. It was decided from the start to minimize moving parts, and the coder is therefore entirely electronic, using transistors, apart from certain essential versions, one a single channel unit for use when the output of only one receiver is being recorded and the other a multi-channel unit which permits the outputs of up to eleven receivers to be recorded on one tape.

Before describing details of the design, a general description of the principle employed will be given. The main function of the equipment is to transform the input voltage (0 to -3V) into five-digit binary form. This means that the five control outputs representing values in 100mV units of 16, 8, 4, 2 and 1 must be formed, each output being in either the 'on' or 'off' state.

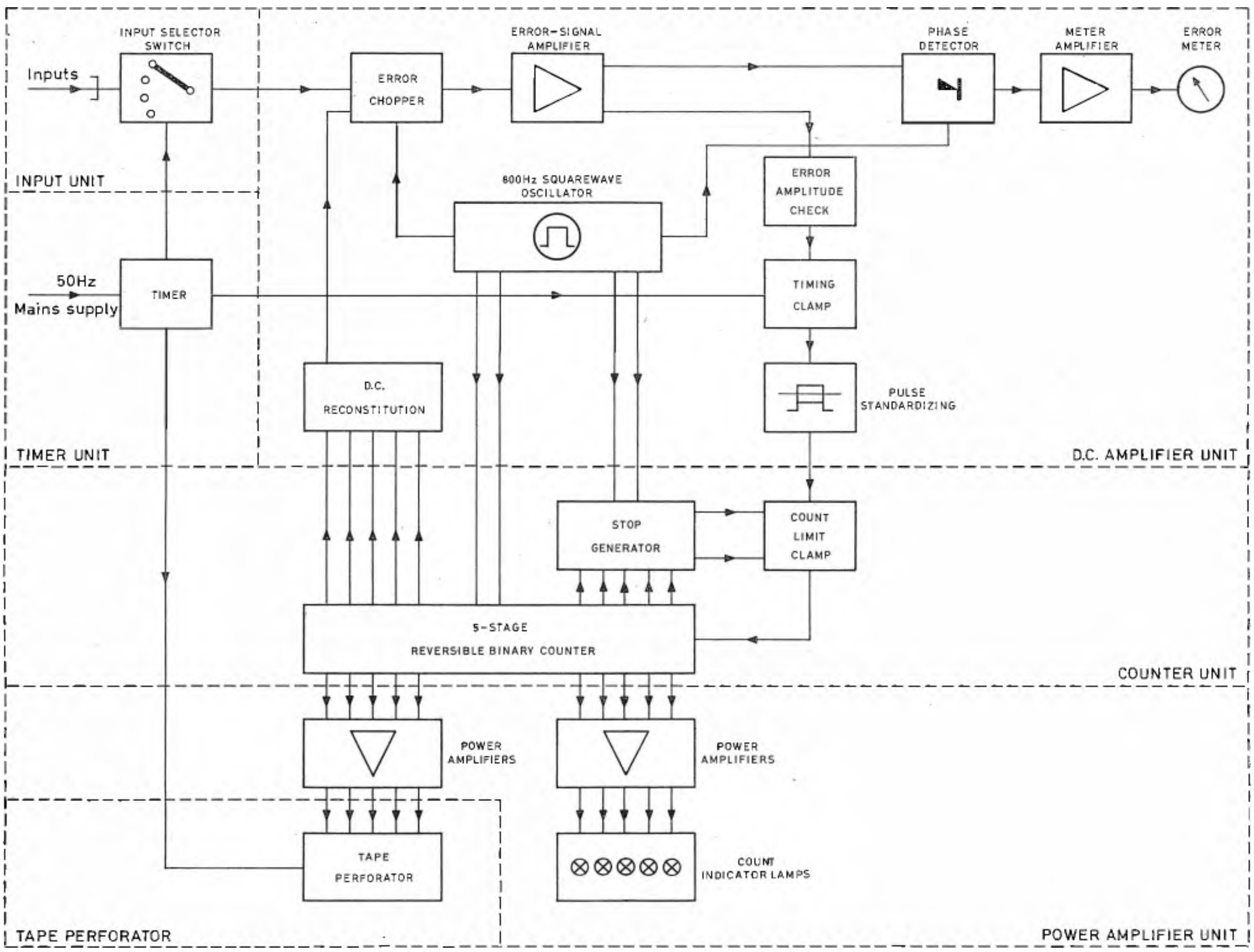


Fig. 2. Arrangement of complete apparatus

components of the tape perforator itself. A prototype equipment has been in use for more than a year and has given good agreement between automatic and hand analysis.

Although the coder was designed primarily for use with field strength measuring receivers it can, of course, be adapted for coding any variant which can be made to produce a voltage in the range of 0 to -3V; for example, a similar circuit is being used in equipment designed to record variations in the frequency of the mains supply.

A photograph of a coder and tape perforator is shown in Fig. 1.

Description of Apparatus

A block schematic of the complete apparatus is given in Fig. 2. The coder consists of six sub-units which plug into a 19in (483mm) bay mounting frame carrying the interconnecting cables. The input unit is made in two

The transformation is achieved with a reversible binary counter by steering the counter to the correct value in the following way. The total value that corresponds to the state of the counter at any moment is formed as a voltage by a reconstitution circuit. This adds currents corresponding to the binary values 1 to 16 according to the state of each of the five stages of the counter, thus giving the reconstituted voltage. This voltage is compared with the input voltage to produce an error signal which is chopped by an 800Hz square wave. Thus the error voltage is converted into a square-wave whose amplitude indicates the magnitude of the error and whose phase relative to the chopping wave determines the sense of the error. This square-wave is amplified and applied (a) to a phase-sensitive detector circuit for displaying the error on a d.c. meter, and (b) to a pulse-forming network which, if the error signal exceeds a preset value, provides positive pulses for the counter unit.

The positive pulses will occur in either the positive or negative half-cycle of the 800Hz reference waveform, dependent on the sign of the error voltage. In operation, the condition of the counter as to whether it counts upwards or downwards is reversed by alternate half-cycles of the reference waveform, so that any input pulse occurring during the positive phase will give an upward count while one occurring during the negative phase will give a downward count. In this manner the count and the reconstituted voltage follow the input voltage and, provided that the input variations are not too rapid, the error is kept to less than one step (100mV). The 800Hz switching is in fact sufficient to ensure that any slowly fading signal passed by a filter at the input to the coder can be followed by the counter.

After a period during which the counter follows the signal the count is temporarily blocked. This occurs at

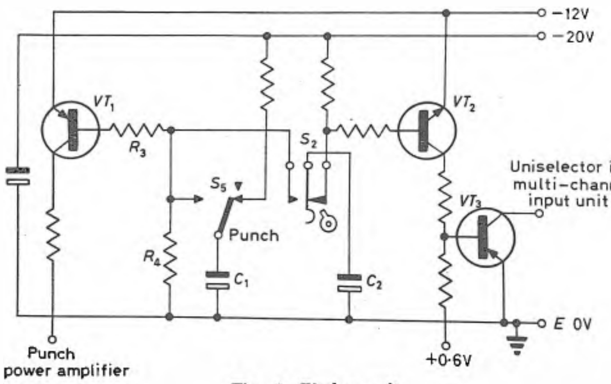


Fig. 3. Timing unit

selected intervals of one, two or three times a minute as determined by the timing cam. The perforator solenoids are set according to the binary count and corresponding holes are then punched in the paper tape. The count conditions of zero and thirty are detected by the 'stop' generator which operates the count-limit clamp so as to prevent the counter from stepping further and giving false counts.

Some of the salient points of each unit will now be discussed.

The Timing Unit

A simplified circuit diagram of the timing unit is given in Fig. 3. The sequence of operations within the coder is governed by the timing unit which contains two cam-operated microswitches actuated by a 1rev/min synchronous motor. Interchangeable cams are provided to give either one, two or three punches per minute. One of the switches operates the timing clamp and permits the signals to pass to the counters for half the total time, after which the clamp is released, the counter blocked, and the code solenoids set. Near the centre of the blocked period the other switch S_2 closes and initiates the operation of the tape perforator and, when it is used, the stepping uniselector in the multi-channel input unit. Before S_2 operates, a large capacitor C_2 charges to −20V. The operation of the switch S_2 connects this capacitor to the base of VT_1 via R_3 and so permits VT_1 to conduct heavily. The output from VT_1 drives the punch solenoid amplifier in the perforator unit. After approximately 0·25sec the capacitor C_2 will have discharged sufficiently to enable VT_1 to cease conduction, thereby releasing the punch solenoid. The capacitor C_2 continues to discharge through R_4. When the switch S_2 releases, the transistor VT_2 is driven into conduction and C_2 commences to charge. VT_2 conducting drives the power amplifier VT_3 into conduction and steps the uniselector in the multi-channel input unit. After approximately 50msec C_2 has recharged to −12V and VT_2 will be cut-off. C_2 continues to charge to −20V before the cycle is repeated.

A manual punch switch S_5 is provided; this operates in a similar manner to S_2 for the punch transistor VT_1 but does not operate the stepping mechanism of the uniselector in the multi-channel input unit.

The Input Units

The input unit is made in two versions. When the output of only one receiver is being recorded the single-channel unit is used; this passes the input directly to the d.c. amplifier unit. The other version is the multi-channel unit, which permits inputs from up to eleven receivers to be recorded on the same tape. A uniselector selects each input in turn, and inserts an 'all-blank' into the code at the end of each sequence. If fewer inputs than eleven are required, then inputs may be paralleled to increase the sampling rate and extra 'all-blanks' inserted as necessary. The extra 'all-blanks' are inserted by connecting the appropriate pins on an internal 12-pin plug.

The uniselector is also used to switch lamps to indicate which input is connected to the coder. A push-button is provided to enable the uniselector to be operated manually.

The D.C. Amplifier Unit

A complete circuit diagram of the d.c. amplifier unit is given in Fig. 4.

The input to the unit passes through a simple low-pass filter R_2C_2 to remove any residual hum or modulation on the signal. The input voltage and the output of the d.c. reconstitution circuit are clamped by VT_1 for alternate half cycles of the 800Hz square-wave oscillator VT_{14} and VT_{15}. Any square-wave produced by the alternate clamping and releasing action of VT_1 is amplified by the three-stage amplifier VT_2 to VT_4. Feedback on the third stage is adjusted by the TRIP LEVEL control R_{21} so that the overall gain of the amplifier may be preset.

One output from the amplifier is clamped to −12V by VT_5 during the clamp phase of the square-wave oscillator. The low-pass filter R_{24}, C_{11}, feeds the resulting mean d.c. level from this clamp to one input of a long-tailed pair amplifier VT_6, VT_7. The other input of the pair is fed from a fixed potential preset by the SET ZERO control R_{33}. A resistor R_{70} between the emitters stabilizes the gain and a variable resistor R_{72} adjusts the balance of the collector loads so as to enable the temperature coefficient of the zero of the amplifier to be set accurately. A meter between the collectors indicates the error between the reconstituted voltage and the input.

The square-wave output from the amplifier VT_2 to VT_4 is also applied to the base of VT_{16} after d.c. restoration by MR_3 so that it is negative with respect to earth potential. The latter transistor has a fixed potential of about −1·5V on its emitter and will conduct only when the peak-to-peak value of the square-wave from the amplifier exceeds 1·5V. The output from VT_{16} collector is applied to the 'pulse standardizer' consisting of a Schmitt trigger pair VT_{18}, VT_{19}. As long as the error signal is sufficient, this provides triggering pulses for the counter via the phase-reversing amplifier VT_{20}. In order to block the counter before recording a sample, VT_{17} is made heavily conducting by the appropriate timing signal and clamps the input to the Schmitt trigger at −3V, thus preventing VT_{18} from conducting.

The transistors VT_8 to VT_{12} are connected one to each of the five binary counter outputs. These transistors will be either cut-off or heavily conducting. When conducting, their base-emitter potentials are less than 100mV. One

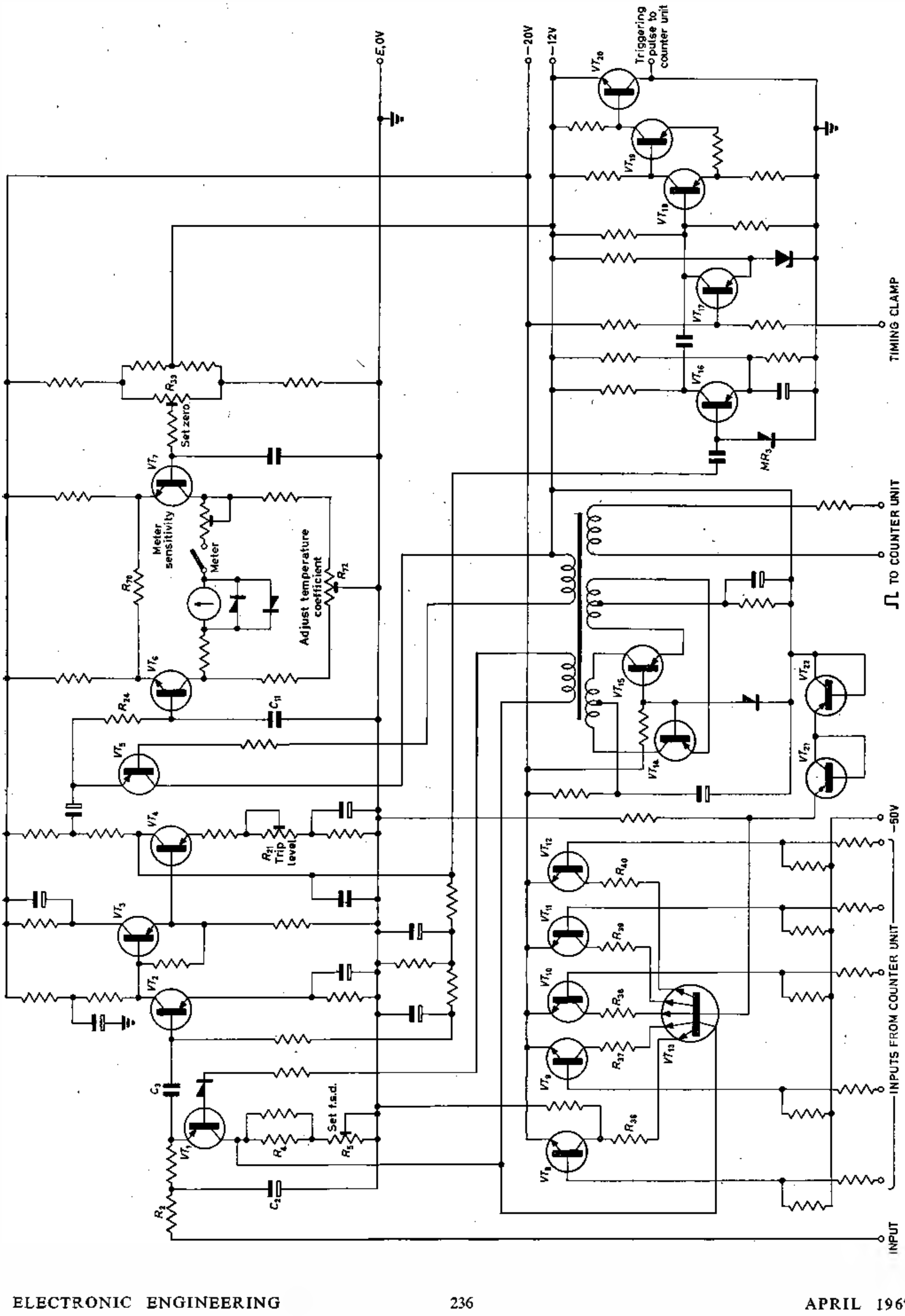


Fig. 4. D.C. amplifier unit

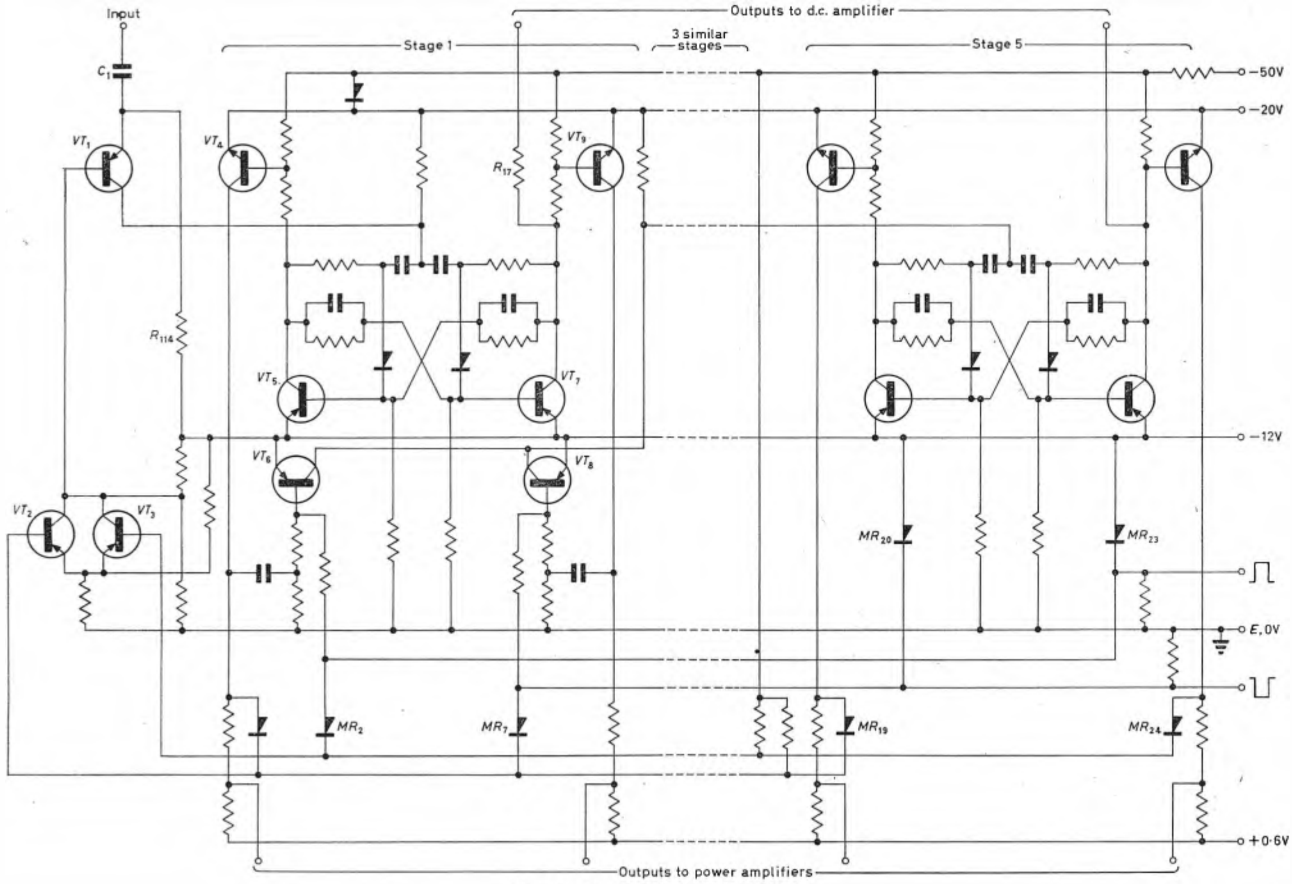


Fig. 5. Simplified counter unit

end of each collector load of these switching transistors is connected to one emitter of a five emitter transistor VT_{13}. The base of this transistor is connected to a fixed potential of approximately −11·5V. Since the five loads R_{36}-R_{40} have resistance values in binary steps, the current passing will also be in binary steps. The collector current of VT_{13} is closely equal to the sum of its emitter currents but is independent of collector voltage, so that the voltage across the collector load R_4, R_5 will be proportional to the count in the binary counter. This voltage is the reconstituted voltage which is balanced against the input to the coder.

Some adjustment of the maximum value of the reconstituted voltage is given by the variable resistor R_5 so that changes of up to $\pm 0{\cdot}25$V in the input corresponding to the maximum count may be made. The change with temperature of the base-to-emitter drop of VT_{13} and emitter-to-collector drop of VT_8 to VT_{12} is compensated by the drop across transistors VT_{21} and VT_{22} connected to operate as low-impedance diodes.

The Counter Unit and Power Amplifier

A simplified circuit diagram of the counter unit is given in Fig. 5. The counter consists of five similar stages of Eccles-Jordan binary pairs, VT_5, VT_7, etc. The input to the counter from the pulse standardizer is differentiated by C_1, R_{114} and applied to the emitter of VT_1. This transistor is normally just cut-off, but the pulse is sufficient to drive VT_1 into conduction and so operate the counter. In order that the counter may be driven either forwards (increasing count) or backwards (decreasing count), as appropriate to reduce the error between the reconstituted voltage and the input to be coded, all four interstage couplings are modified every half-cycle of the reference 800Hz switching signal. This function is performed by

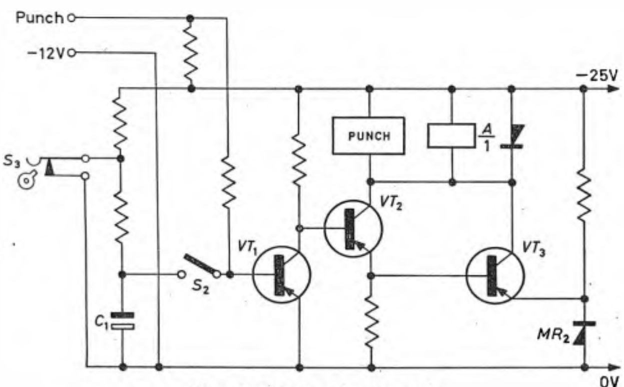


Fig. 6. Tape perforator unit

passing the 'carry one' count from one stage to the next via pairs of transistors such as VT_6, VT_8 shown for the first stage in Fig. 5. One member of the pair is held just cut-off and the other well beyond cut-off by the 800Hz square-wave; their conditions are therefore reversed each half-cycle. A trigger pulse is passed on only through the transistor that is just cut-off and, since the transistors are connected to opposite sides of the binary stage, the count is passed on for a 0 to 1 transition for the first half-cycle and for a 1 to 0 transition for the second half-cycle. This means the direction of count is in the reverse direction for alternate half-cycles as required by the principle of operation already explained. The output winding of the 800Hz oscillator is floating, but diodes MR_{20} and MR_{23} are used to self-bias the square-wave to the correct potential, thus eliminating the effect of any changes in amplitude of the oscillation.

When the counter has reached either a maximum or minimum count, further changes in the same direction must be prevented or an indeterminate count would be obtained. These limiting counts are detected by the diodes MR_{19}, MR_{24}, etc., and when the limit is reached and the square-wave, as detected by MR_2, MR_7, is in the appro-

solenoids are gated so that the solenoids are not operated during the time that the counters are running, thus preventing undue wear of the mechanical linkages.

A conventional power supply unit provides an unstabilized output of −50V and stabilized outputs of −20V,

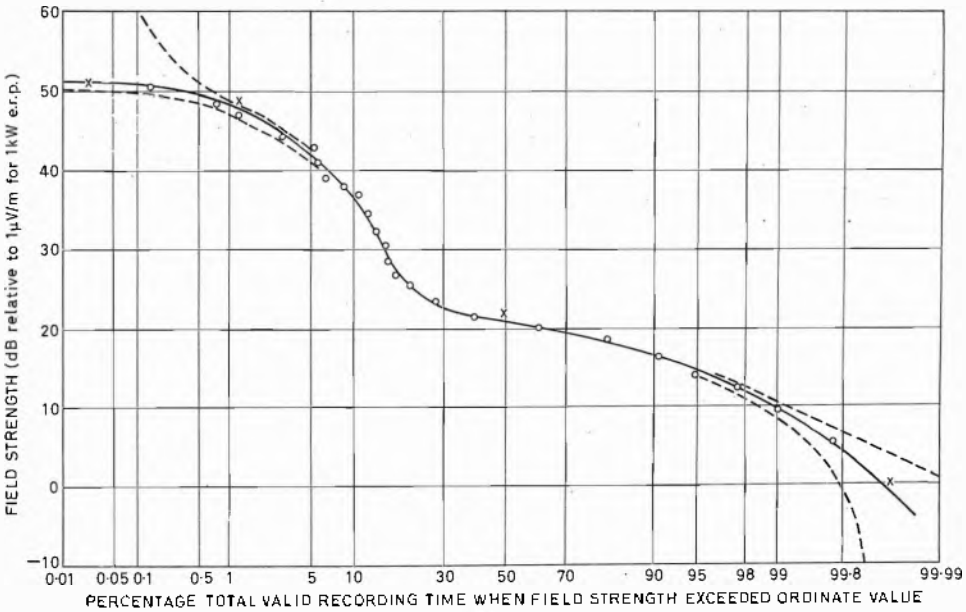


Fig. 7. Comparison of coder and recorder chart analysis

° *Coder tape analysis* × *Recorder chart analysis* ---- *95% confidence limits*

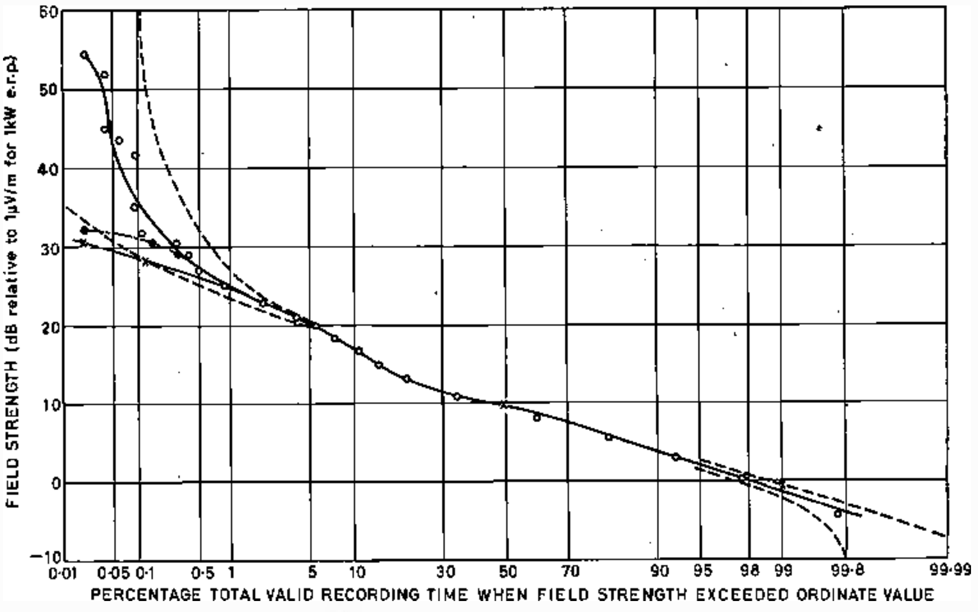


Fig. 8. Comparison of coder and recorder chart analysis

° *Coder tape analysis* × *Recorder chart analysis* • *Corrected coder tape analysis* ---- *95% confidence limits*

priate direction, either VT_2 or VT_3 will conduct. This will take VT_1 well beyond cut-off and prevent any further count in the same direction.

Each of the counter stages has three outputs to other units. The first drives the d.c. reconstitution circuit via R_{17}, etc., while the other two have output amplifiers VT_4, VT_9, etc. These outputs drive power amplifiers for operating the code solenoids in the perforator unit and indicator lamps to show the state of the count.

The power supplies to the amplifiers driving the code

−12V and +0·6V. A protective device cuts off the power supply in the event of a severe overload.

The Tape Perforator Unit

This unit contains a modified commercial tape perforator together with power supplies and a spooling motor. The principal modifications to the perforator are a new chad shute and the substitution of a d.c. operated rotary solenoid for the a.c. solenoid. The unit, a simplified circuit of which is shown in Fig. 6, contains a three-stage transistor amplifier VT_1 to VT_3 to operate the punch solenoid. The first transistor VT_1 is normally conducting heavily but is cut-off by an incoming punch pulse. The low collector potential of VT_1 together with the 0·6V drop across the diode MR_2 maintains VT_2 and VT_3 cut-off except during the punch pulse when these transistors will conduct. The transistors VT_2 and VT_3 are connected as a grounded emitter Darlington pair and on conduction will pass a large (4A) current through the solenoid. The power supply is a simple bridge rectifier circuit giving about 25V off load. This voltage falls to about half during the punch pulse. The code solenoids are directly connected to the coder outputs.

In order to 'run through' paper tape, twin push-buttons are provided. One of these switches removes the supplies from the code solenoids and the other, S_2, connects a discharged capacitor C_1 across the base of VT_1, causing the punch solenoid to operate. During the punch operation the cam-driven switch S_3 operates and allows C_1 to charge, thereby releasing the punch. This process is repeated continuously while the buttons are depressed.

The mains-operated tape-winder has its power supplies connected by relay A during the punch pulse only.

Results

The coder translates into binary form a 0 to −3V input giving thirty steps of 100mV. The centre point of these steps is within ±2mV of the correct value and will not change by more than ±2mV over a temperature range of −10°C to +45°C. The stepping rate of the counter is 800 steps per second giving a maximum balance time of less than 40msec. The following rate on a varying signal is limited by the low-pass filter on the coder input to a few cycles per second, dependent on the source impedance.

Some results of experiments using the prototype coder equipment are given in Figs. 7, 8 and 9. It should be remembered that a sampling system is being used and in consequence sufficient samples are required in each level before that level can be stated with any confidence. As a guide the 95 per cent confidence limits, i.e., the points where there is a 95 per cent probability that the true value lies within the given boundaries, are shown for the extremities of Figs. 7-9.

Fig. 7 is an example of good agreement between the results of the coder tape and the recorder chart analysis. In Fig. 8 the coder results show some exceptionally high signals compared with the chart analysis. These may be due to some form of impulsive interference which can be recorded by the coder but not by the recording meter which has a slower response time.

In Fig. 8 a few high level counts out of a total population in excess of 5 000 are responsible for the shape of the curve at low percentage times. If the high level counts are ignored and the percentage recording time recalculated then the chain dotted curve of Fig. 8 is obtained. This curve is in reasonable agreement with the recorder chart analysis.

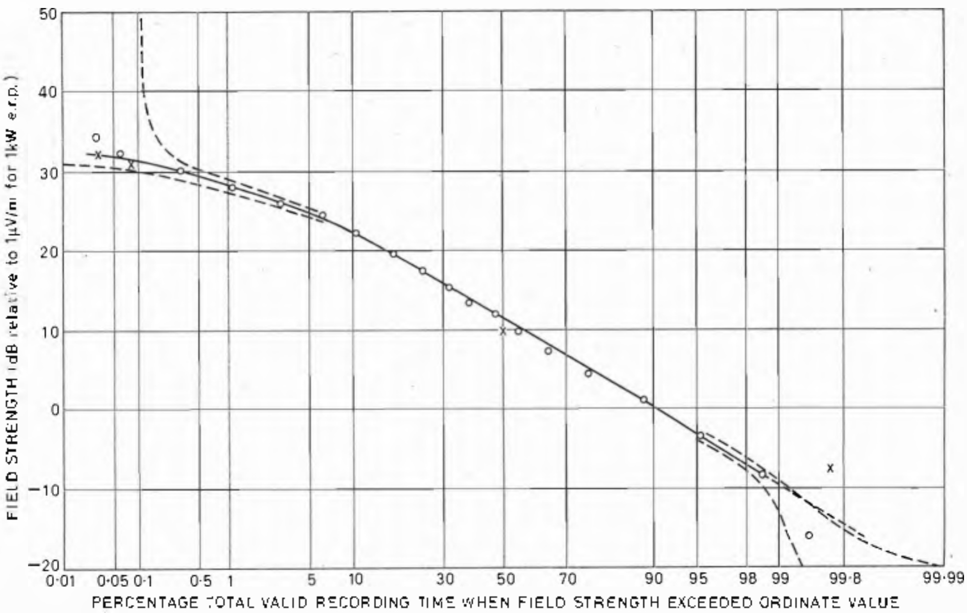


Fig. 9. Comparison of coder and recorder chart analysis

° *Coder tape analysis* × *Recorder chart analysis* ---- *95% confidence limits*

Fig. 9 is a further comparison of results, but in this case errors occur at low fields where the coder value appears to be too low and the recorder chart value too high. The low coder value can be accounted for by either breaks in transmission or low power conditions at the transmitter. The high field strength from the recorder chart is probably due to pen or pivot friction.

Apart from the few anomalies discussed above, the results of the field strength measurements using perforated tape are in good agreement with the recorder chart data. The differences indicate that care is required in interpreting the results from both the coder and the recorder chart especially where few counts at either high or low levels are involved. The man-hours required for analysis have been very considerably reduced by the introduction of computer techniques.

Acknowledgments

The authors wish to acknowledge the assistance of all their colleagues in the BBC Research Department who helped in the construction of the equipment, and to thank J. W. Stark for providing the comparisons of manual and computer analyses of the records. They are also indebted to the Director of Engineering of the BBC for permission to publish this article.

An Analyser for Use in Human Electromyography

By P. Fitch*

An instrument for the automatic analysis of electromyographic signals is described. Intended for use in clinical diagnosis it produces two sets of serial pulses representing respectively the occurrence of peaks in the waveform, and their amplitude. The method of analysis is such that all variations of the signal voltage are processed without reference to its absolute value.

(Voir page 275 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

THE electromyogram (e.m.g.) which is recorded from a coaxial needle electrode in human muscle during voluntary effort, is widely used in the diagnosis of diseases of the muscle and nerves. It consists of a grossly irregular series of waves occupying a bandwidth of about 10Hz to 5kHz and having amplitudes of up to 15mV. When the activity is viewed on a c.r.t. the 'spikiness' and amplitude of the waves are two features of diagnostic value. The assessment of these features is at present mainly subjective. In order to give a quantitative description of the e.m.g. a photographic technique was developed[1]. This method was extremely laborious and the present apparatus was designed to carry out similar measurements automatically and in real time. Essential features of the method are that all changes of potential having an amplitude greater than a chosen unit value are measured, and the total of all such measurements during a given time interval is accumulated; during the same period a count is made of 'turns' in the waveform (a 'turn' being defined as the occurrence of two successive increments of signal amplitude whose slopes are of opposite sign and whose amplitudes are not less than the unit value). It will be seen that the 'spikiness' of the waveform is not in this system defined by the number of 'zero crossings.'

In the apparatus described the unit value for all signal measurements is $100\mu V$ at the electrode, and to measure signal amplitude the apparatus delivers a pulse to a counter after each signal increment of $100\mu V$ without regard to the sign of the incremental slope or to the absolute signal potential at that time, see Fig. 1. Pulses marking the occurrence of 'turns' in the waveform are delivered to a second counter whenever two successive signal increments of $100\mu V$ or more have slopes of opposite sign. It is self-evident that the maximum rate of occurrence of amplitude registering pulses is, for any given waveform, inversely proportional to the size of the units in which it is measured. Thus a unit of amplitude of $100\mu V$ was chosen (after preliminary consideration of the technique employed), as being the smallest practical unit, taking into account the amplitude and bandwidth of the signal in relation to the limitations imposed by components and measuring equipment then available to the author.

General Description

The arrangement of the system adopted for amplitude measurement is shown in Fig. 2. Muscle action potentials from a needle electrode are amplified in a conventional way, and the signal appearing at the monitor c.r.o. is applied to C_2 via the drive amplifier which has a low output impedance. The scale of the signal at this point is adjusted so that a $100\mu V$ signal at the electrode is represented by 1V. The signal appearing across R_2 is applied via the cathode-follower to the comparator circuits, which respond to signals of 1V positive or negative to the zero line. The activation of either comparator causes the monostable circuit to switch the clamp circuit into its conducting state for approximately 400nsec, bringing the junction of C_2 and R_2 to zero level. Thus the next increment of the signal at the cathode-follower input starts at zero potential and activates one or other comparator on reaching a potential of 1V thereby causing the clamp to operate again. Each operation of the clamp is counted

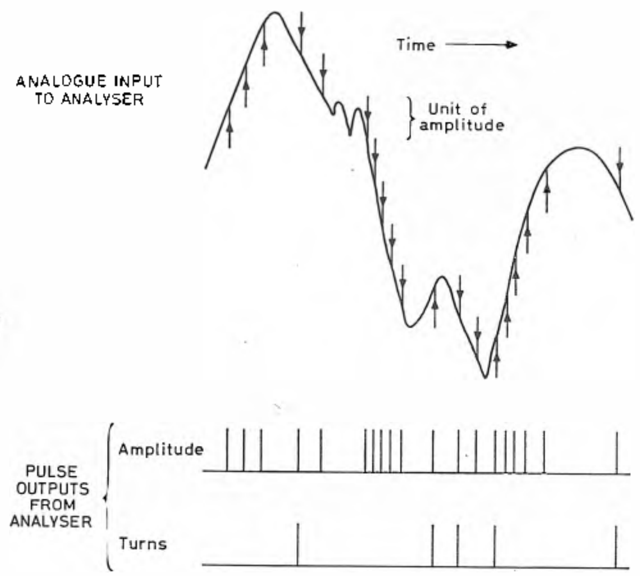


Fig. 1. Diagram illustrating the analysis of a typical waveform, according to the criteria described in the text, and showing the relationship between the analogue input and the analyser outputs

by the amplitude counter, and the total represents the sum of the amplitudes of all the excursions of the signal potential in units of $100\mu V$, during the time chosen.

Although the addition of suitable logic units fed from the comparators would make it possible to deliver to a second counter one pulse for each turn in the waveform as previously defined, the detection of turns is currently achieved by a different technique, because the comparators may fail to detect some signals having an amplitude of nearly $200\mu V$. Consider such a signal as it appears at the junction of C_2 and R_2 in Fig. 2, where each $100\mu V$ increment is represented by 1V. If, for example, after a clamp operation there now follows a signal which reaches plus 0·9V before turning in a negative direction, the signal will have to traverse 1·9V if it is to operate the negative comparator; so that if after the signal reversal at plus 0·9V there follows a negative going in-

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crement of only say 1·8V, neither comparator will have operated although this part of the signal exceeded the unit size of 1·0V. The modified technique evolved to eliminate this kind of ambiguity is shown in Fig. 2. The analogue signal at the output of the drive amplifier is applied to the capacitor C_1 which is connected to the comparator circuits via the cathode-follower. The comparators are set to respond at 1·0V positive or negative, and the activation of either sets the bistable into a corresponding stable state, so that as the positive comparator is activated, the clamp connected to the diode MR_1 is switched into its conducting state, bringing the junction of C_1 and R_1 to zero level. Similarly the clamp connected to MR_2 will close when the negative comparator is activated. The closing of the positive clamp prevents further positive going increments from appearing at the comparator input, but (due to the presence of the diodes MR_1 and MR_2) a negative increment exceeding 1·0V can operate the second comparator circuit so that the condition of the bistable circuit is reversed, opening the clamp previously closed and closing the other. A similar sequence will occur for signals of opposite polarity; thus each operation of a comparator represents a turn in the voltage waveform. The comparator outputs are fed via the OR gate to the turns counter. The turns counter and the amplitude counter are controlled by common time pulses so that both count for the same interval of time (usually in the range of 1 to 5sec). However, the possibility still exists that the amplitude detection system will disregard some low amplitude portions of the signal waveform. This shortcoming can be overcome by arranging that the monostable circuit driving the clamp is triggered additionally by each turn as detected by the turns detector, so that subsequent amplitude measurements are constrained to begin from each turn in the waveform. Closing the link between the turns pulse output and the lower OR gate (Fig. 2) completes the schematic diagram of the combined system.

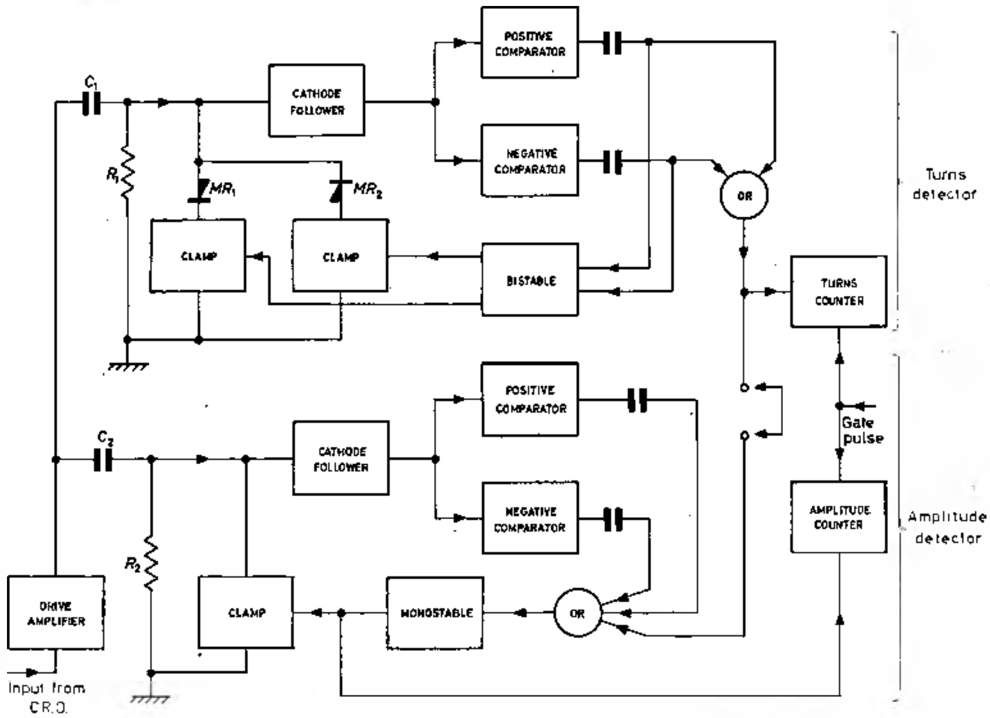

Fig. 2. Arrangement of complete analyser

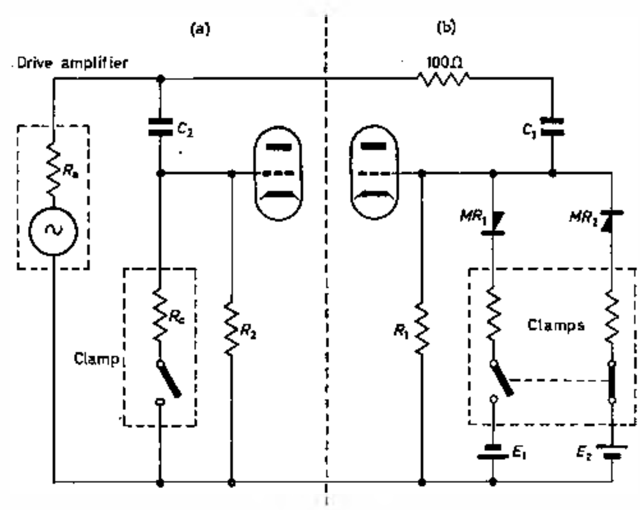

Fig. 3. (a) Simplified equivalent circuit of clamp portion of amplitude detector
(b) Input elements of turns detection circuit

Practical Design Considerations

Fig. 3(a) shows a simplified equivalent circuit of the clamp portion of the amplitude detector. The scale of the signal at the output of the drive amplifier is constrained as follows. Firstly the dynamic range is set by the ratio of the unit signal of 100μV to the maximum signal normally occurring; existing photographic e.m.g. records indicated that provision for maximum signals of 5mV peak-to-peak would usually be sufficient (for the type of investigation envisaged) i.e., a dynamic range of 50 : 1. As previously indicated, a scale giving a unit signal of 1V at this point was chosen in spite of the resulting maximum signal of 50V, a smaller unit signal would simplify the amplifier requirements, but the offset voltage of the clamp transistors would introduce proportionately larger errors in the amplitude measurement. The duration of the clamp closure is also a compromise between conflicting requirements. The limiting duration of the clamp closure time may be approximated as follows. A sinusoidal signal having a frequency of 5kHz and an amplitude of 50V peak-to-peak will cause the maximum rate of clamp operation to occur, since both signal frequency and amplitude are maximal.

This signal can be shown to have a maximum rate of change of signal potential with respect to time of $7{\cdot}85 \times 10^5$V/sec, thus the interval between one volt increments tends to a minimum of

$$1/7{\cdot}85 \times 10^5\text{sec} = 1{\cdot}28\mu\text{sec}$$

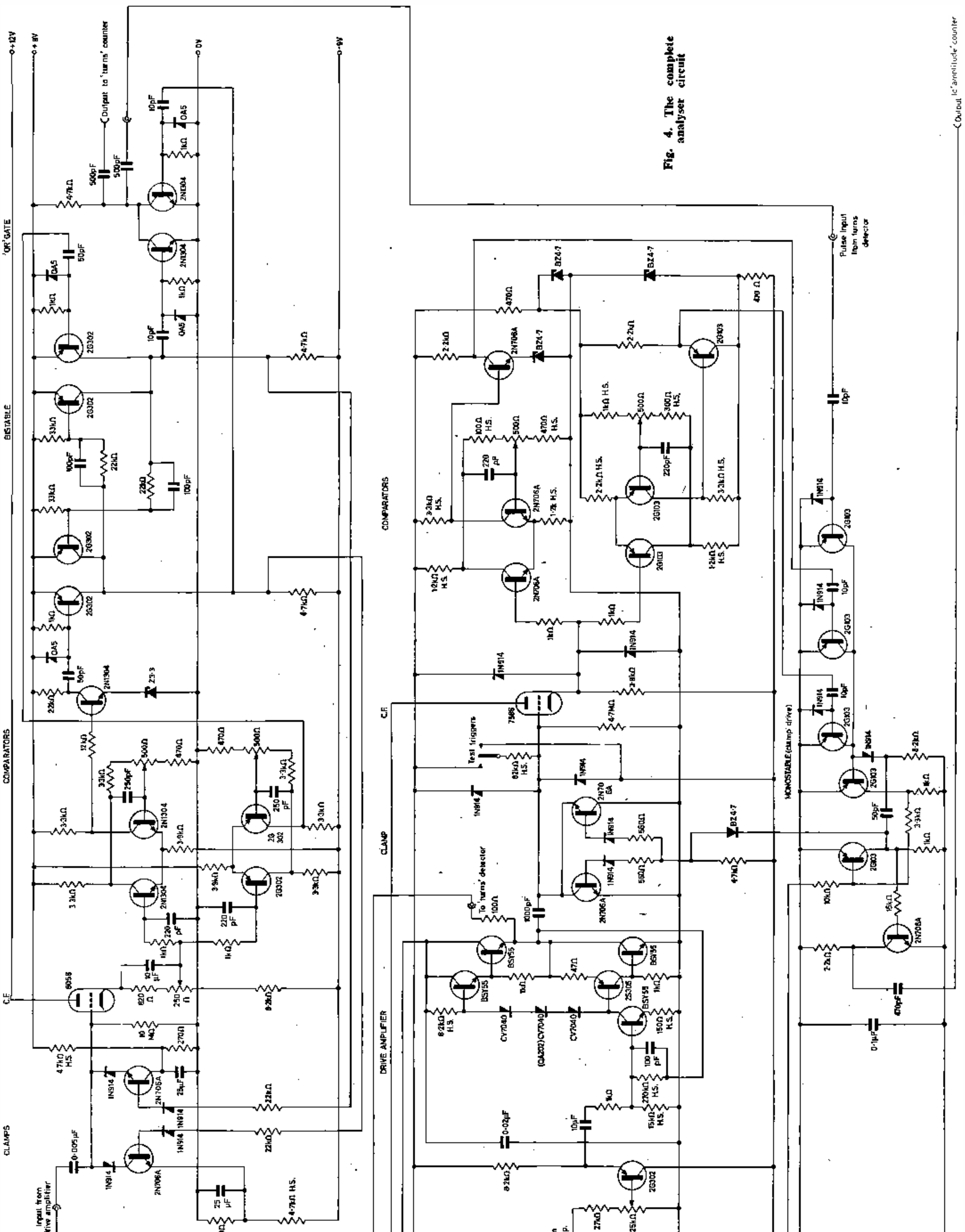

Fig. 4. The complete analyser circuit

Since the occurrence of an amplitude increment of unit size is marked by the clamp closing, the closure time itself should ideally be much less than this. The lower limit of the clamp closure time is set in practice by the need to discharge C_2 adequately through $R_a + R_c$, where C_2 has a minimum capacitance defined both by the desired l.f. response and the highest practical input impedance at the cathode-follower. The lower limit for the value of $R_a + R_c$ is set by the practical limitations of the amplifier and clamp circuits. A clamp closure time of approximately 400nsec is used in the apparatus constructed, which limits C_2 to a value of 1000pF since R_a and R_c have values of about 6Ω and 45Ω respectively, so that the clamp closure time has a duration of about $8\,C_2(R_a + R_c)$ seconds. The value of R_1 is 4·7MΩ giving an l.f. 3dB point of around 30Hz.

Fig. 3(b) shows the input elements of the turns detection circuit. Since a 5kHz signal has 100μsec intervals between turns, the time constraints on this circuit are less severe than the case of the amplitude detector. Hence C_1 has a value of 5 000pF, and the drive amplifier impedance has been raised to 100Ω in order to limit the peak current drawn from the amplifier. The forward voltages across the diodes MR_1 and MR_2 are backed off by E_1 and E_2 at the cost of reducing the reverse impedance to about 1MΩ. Because C_1 has a value of 5000pF the l.f. 3dB point is the same as for the amplitude detector.

The circuit diagram of the complete analyser is shown in Fig. 4. Adjustable Schmitt triggers perform the comparator functions and their operate levels are set by observing the signal input to the cathode-followers with a c.r.o. It is hoped to develop a more convenient calibration system in the near future; however, the stability of these circuits is such that frequent adjustment has not been necessary. Observations of the amplitude count totalled by sinusoidal calibration signals in relation to the number of turns totalled in the same period, show that the mean amplitude indicated by this technique has an error at maximum amplitude of up to 5 per cent for frequencies below 1kHz, increasing to 20 per cent at 5kHz. Since the measured error at 5kHz is of the order of magnitude expected from the clamp closure time (400nsec), an increase in the accuracy of amplitude measurements at the higher frequencies is dependent upon reducing the clamp closure time still further. This will involve overcoming the present limitations of the amplifier and clamp impedances.

Conclusions

The apparatus has been usefully applied to studies in quantitative electromyography[2]. The availability of output pulses relating both to amplitude and 'turns' in the waveform has opened up a number of possibilities for the further analysis of various aspects of the signal waveform. An important property of the apparatus is that the occurrence of transient signals with an amplitude exceeding the range of the instrument will result only in the loss of information and will not generate spurious information due to overload effects.

Acknowledgments

Thanks are due to the following, Mr. J. H. Aird of Elliot Automation Ltd, for suggesting the clamping technique used in the amplitude measuring section.

The Muscular Dystrophy Group of Great Britain for a grant towards the cost of components and apparatus.

Mr. H. B. Morton of the National Hospital, Queen Square, W.C.1, for many helpful suggestions and criticisms.

The Decca Record Co. Ltd for producing the gramophone recordings of e.m.g. signals which were used to test the apparatus.

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Automatic Checking Equipment for Aircraft

T.R.A.C.E. (Tape-controlled Recording Automatic Check-out Equipment) is now being extensively used throughout the world's airlines. British Overseas Airways Corporation has had one equipment in use for over two years and has recently ordered a second; two equipments are on order for British European Airways; a second equipment is on order for the R.A.F. and equipments are also on order for Alitalia and Pan American.

T.R.A.C.E. 600 is the basic series of automatic test equipments now being produced in quantity by Hawker Siddeley Dynamics Ltd for use by the world's civil airlines, military air transport organizations and aircraft manufacturers. It is used to test units which have been removed from an aircraft, and can either act as a confidence checker or as a diagnostic tool.

T.R.A.C.E. is a universal automatic test system which applies to the unit under test all the necessary signal and power sources and subsequently carries out complete dynamic tests, including the measurement and recording of voltages, currents, frequency, resistance, phase angle and synchro angle, etc. The unit to be tested is connected to T.R.A.C.E. by a special purpose coupling unit which simulates the electrical environment normally experienced in the aircraft. Comprehensive self-check features are incorporated to ensure complete validation of the tests carried out.

The basic unit of the T.R.A.C.E. 600 series is designed to test primarily the components of the automatic flight control systems of today's large jet transport aircraft. In addition it can also test comparable electrical/electronic items such as anti-skid computers, generator control panels, cabin address amplifiers, etc. With the simple plug-in addition of optional extras its testing capability can be extended to include pneumatic components, e.g. air data computers, flight instruments; radio components, e.g. distance measuring equipment, doppler and communication systems. Further "add-on" extras include an arithmetic unit to perform mathematical operations on the test results and a tape punch to obtain the results in a convenient form for off-line analysis by standard data processing computers.

The T.R.A.C.E. 600 series is an improved version of the T.R.A.C.E. III which has been operated satisfactorily by the British Overseas Airways Corporation as a filter tester for their VC.10 and Boeing 707 aircraft since October 1964. By the use of T.R.A.C.E. during this period B.O.A.C. have reduced the input of units to their repair workshop by over 80 per cent. As a consequence they have recently placed an order for a second T.R.A.C.E. equipment to be installed at London Airport in May 1967.

The BEA T.R.A.C.E. 600 will be installed at BEA's maintenance area at London Heathrow airport before the end of the year.

This equipment will be used to ground-test units of the Smith's Flight Control System which will enable BEA's fleet of Hawker Siddeley Tridents to land automatically in all weathers. It will also be used to test the flight-system installed in the new One-Eleven 500 jets.

Of the two units one will be stationed in the Central Area at London Airport, Heathrow, to make possible on the spot filtering-out of faulty components. The defective units will then be passed for rectification to the Engineering Base, complete with an automatic print-out identifying the fault, and later be re-tested after overhaul on the Base unit.

Redundancy in Digital Systems

By F. A. Inskip*

A number of methods of using redundancy in digital systems is described. Numerical values of reliability are given for a small system of the order of size typical of a small data handling system. Two values of reliability are given; one using the best values of component reliability available at the present time and the other using an order lower.

(Voir page 276 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

THE growth in complexity of electronic equipment together with the demand for high reliability has made the cost of proving the required component reliability too high to be acceptable. The only practicable way out of this impasse is the use of some form of redundancy so that the equipment will keep working if some component failures occur. In this article four types of redundancy, two of the Moore-Shannon type, one suggested by von Neumann and one by Tryon are compared. The relative increases in cost, power consumption and reliability are given for a small system.

TABLE 1
Probability of Failure of System

PERMISSIBLE NUMBER OF FAILURES N	$N=1$	$N=2$	$N=3$	$N=4$
0	Q	$2Q-Q^2$	$3Q-3Q^2+Q^3$	$4Q-6Q^2+4Q^3-Q^4$
1		Q^2	$3Q^2-2Q^3$	$6Q^2-8Q^3+3Q^4$
2			Q^3	$4Q^3-3Q^4$
3				Q^4

Possible Improvement in Reliability by the Use of Redundancy

The types of redundant systems considered here are those in which each unit of a series system is replaced by two or more similar units so connected that the system works if one or more of each set of units fail.

Let each of the original units be replaced by N units each of reliability R, and unreliability Q where $Q = 1-R$.

Expanding:

$$(R + Q)^N = R^N + NR^{N-1}Q \;\ldots. + Q^N \quad \ldots\ldots \quad (1)$$

the first term on the right is the probability that no units have failed, the second one unit, and so on, the last term being the probability that all units have failed[1]. For practical purposes, N is limited to a fairly small value by increases in cost, weight, and power consumption. Table 1 shows the possible improvement for values of $N \leqslant 4$, assuming that each of the redundant units has the same reliability as the original unit.

Possible Ways of Employing Redundancy

Component Redundancy

With any component, for example a capacitor, which can fail open- or short-ciruit or by change of value, the minimum number which can be used in a redundant set, to protect against one failure, is four connected as in Fig. 1. In order to cover both types of catastrophic failure the circuit must be capable of working with a tolerance of 50 per cent in the value of capacitance if one failure is allowed. Appendix A gives the reliability for the cases of two or more failures. The tolerance must now be increased to 60 per cent.

* *Royal Aircraft Establishment.*

Very few circuits can be made to work with these tolerances. The arrangement is, however, practical in such cases as decoupling or smoothing circuits.

A similar arrangement can be used for diodes if the variation in forward voltage drop is permissible.

Redundancy at Function Level

In digital circuits the system is built up from a number of simple circuits such as AND, OR, NAND, NOR gates. It is possible to replace one gate by a set of gates so that the set still functions if a failure occurs. An alternative system is to triplicate a gate and add a voting element. It can be shown that voting should take place at points separated by more than one gate for optimum results.

Practical systems not employing separate voting circuits are limited to arrangements of three or four gates replacing a single gate.

Triple systems are discussed in Appendix B where it is shown that the reliability is worse than that of a non-redundant system. A system using four gates is shown later to give a large improvement in reliability.

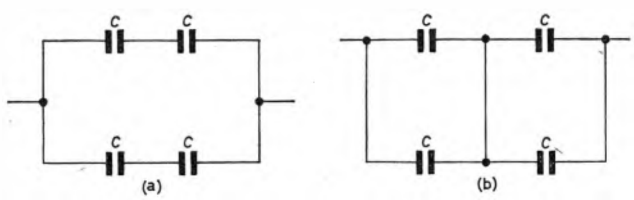
Redundancy at Higher Levels

The use of majority voting is discussed in this article. Another possible system is 'stand-by redundancy', in which a number of complete sub units or equipments are connected to a switch which changes over to a good unit when a unit fails. This method requires some means of knowing when a unit has failed, and automatic failure detection can be very complicated when the equipment is handling unknown information.

Practical Redundant Circuits

Choice of System

Since it was considered desirable to be able to quote numerical values for the improvement possible by using



FAILURE	EFFECTIVE CAPACITANCE ($C=1$) a	b		PROBABILITY OF WORKING a	b
None	1	1		1	1
1 s.c.	$^3/_2$	2		1	1
1 o.c.	$\frac{1}{2}$	$\frac{2}{3}$		1	1
2 s.c.	2	2	(Set still good)	$\frac{2}{3}$	$\frac{1}{3}$
2 o.c.	$\frac{1}{3}$	$\frac{1}{3}$	(Set still good)	$\frac{1}{3}$	$\frac{2}{3}$
1 o.c. +1 s.c.	1 or $\frac{1}{2}$	2 or 1		1	1
2 o.c. +1 s.c.	1	1	(Set still good)	$\frac{1}{3}$	$\frac{2}{3}$
2 s.c. +1 o.c.	1	1	(Set still good)	$\frac{2}{3}$	$\frac{1}{3}$
3 o.c.				0	0
3 s.c.				0	0

Fig. 1. Component redundancy

redundancy it was necessary to choose the most general type of circuit. As any digital circuit can be built up from simple NOR gates this was chosen as the basic circuit unit. A series chain of 48 gates was chosen as this is fairly typical of a small data handling system.

NOR Gate

The circuit and logic systems are shown in Fig. 2.

The best estimate for the reliability of the resistors, assuming considerable derating and a low stress environment is a failure rate less than 10^{-8} per hour. Rates as low as $3{\cdot}10^{-10}$ have been obtained in some cases.

Failure rates for transistors are more difficult to estimate, but a failure rate of 10^{-7} per hour can probably be maintained on selected and aged silicon planar devices operating under low stress conditions. Results are also given for a failure rate of 10^{-6} per hour, because this is a much less costly failure rate to prove.

Component Redundancy

If the collector load resistor is quadruplicated (as in Fig. 1) the power consumption of the circuit is increased several times and the circuit design becomes much more difficult.

Fig. 3(a) shows the basic NOR gate and Figs. 3(b) and 3(c) two ways of connecting redundant transistors. Since, for these circuits, each transistor must be capable of passing sufficient current to bottom, the total drive (base) current is four times that of the basic gate, thus the available fan-out is reduced four times. Since a fan-out of eight is typical for the basic gate the redundant gate is reduced to a fan-out of two, which limits the usefulness of the circuit considerably.

The reliability of the circuit of Fig. 3(b) is calculated as follows.

The table in Fig. 3(b) gives the probability of working for various fault conditions. The reliability is then calculated assuming (a) that the resistor is perfect or is quadruplicated, (b) the reliability of the resistor is 0·9999.

For transistors of very high reliability operating at very low power levels the most probable causes of failure are loss of gain, open-circuit bonds causing an open-circuit failure, and high leakage current causing a short-circuit failure. A ratio of three open-circuit failures to one short-circuit failure is probably realistic.

Let R_g = probability of the gate working
R = probability of a transistor working
Q = probability of a transistor failing
Q_o = probability of a transistor failing open-circuit
Q_s = probability of a transistor failing short-circuit
λ = failure rate of transistor
t = time

Then:

$$R + Q = 1,$$
$$Q = Q_o + Q_s.$$

From equation (1) and table of Fig. 3(b):

$$R_g = R^4 + 4R^3Q + 6R^2(\tfrac{1}{3}Q_o^2 + 2Q_oQ_s + \tfrac{2}{3}Q_s^2) + 4R(Q_o^2Q_s + 2Q_oQ_s^2)$$
$$= 1 - 4Q_o^2 - 2Q_s^2 + 4Q_o^3 - Q_o^4 + Q_{s4}$$

If

$\lambda = 10^{-7}$ then $R = 0{\cdot}999$ and $R_g = 0{\cdot}999998$
$t = 10^4$ $\quad Q_o = 0{\cdot}00075$
$\quad Q_s = 0{\cdot}00025$

The reliability of a two-input gate R_{g2} is given by:

$$R_{g2} = R_g^2$$

The reliability of a chain of 48 two-input gates R_T is given by:

$$R_T = R_{g2}^{48}$$

Similarly it can be shown that for the circuit of Fig. 2(c):

$$R_g = 1 - 2Q_o^2 - 4Q_s^2 + 4Q_s^3 + Q_o^4 - Q_s^4 \quad \ldots\ldots (3)$$

The reliabilities were calculated using these equations assuming (a) that the collector load resistor does not fail or is quadded, and (b) that a single resistor with a failure rate of 10^{-8} per hour is used.

The results are given in Table 2 for the chain of 48 gates.

Gate Redundancy

It is possible to connect gates in series-parallel combinations, similar to those used for components in Fig. 1. Two possible arrangements are shown in Figs. 4 and 5. In each circuit the output resistor is common to the four gates.

The analysis of reliability in terms of transistor failures

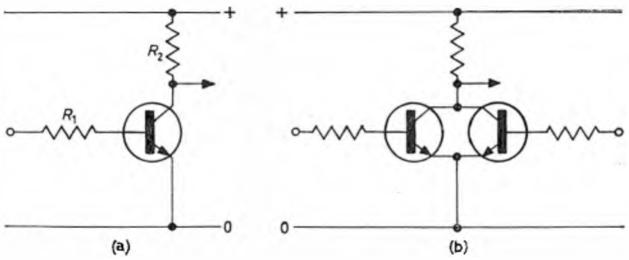
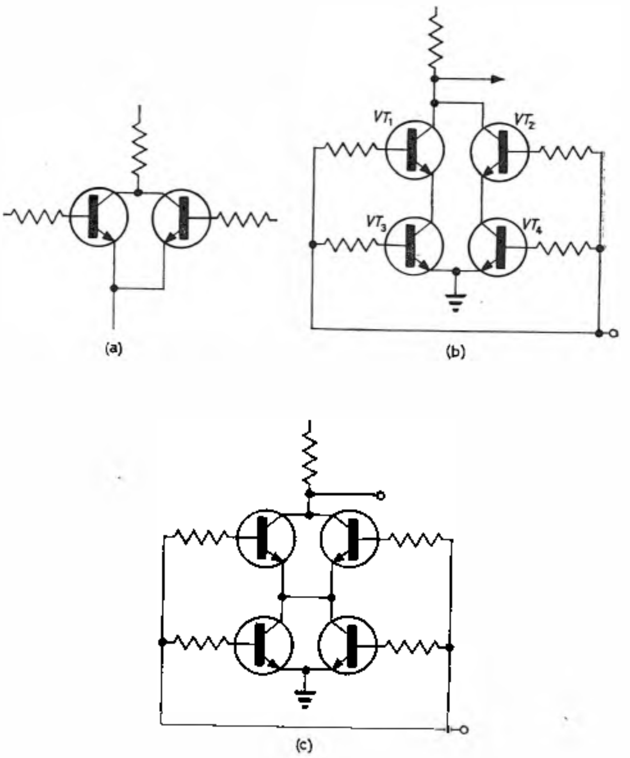


Fig. 2. NOR gates
Logic 0 is 0 volts; 1 is positive voltage

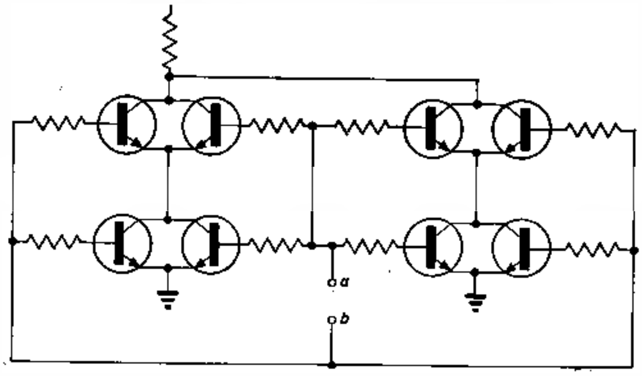


FAULT	PROBABILITY OF WORKING	PROBABILITY OF WORKING
1 s.c.	1	1
1 o.c.	1	1
2 s.c.	$\frac{2}{3}$	$\frac{1}{3}$
2 o.c.	$\frac{1}{3}$	$\frac{2}{3}$
1 o.c. + 1 s.c.	1	1
3 o.c.	0	0
3 s.c.	0	0
2 o.c. + 1 s.c.	$\frac{1}{3}$	$\frac{2}{3}$
2 s.c. + 1 o.c.	$\frac{2}{3}$	$\frac{1}{3}$

Fig. 3. NOR gate with redundant transistors

TABLE 2
Reliability of System for 10^4 Hours

FAILURE RATE		Fig. 3(b)	Fig. 3(c)
Transistor	*Resistor*		
10^{-7}	0	0·99977	0·99987
10^{-6}	0	0·97761	0·98689
10^{-7}	10^{-8}	0·99498	0·99508
10^{-6}	10^{-8}	0·97293	0·98217

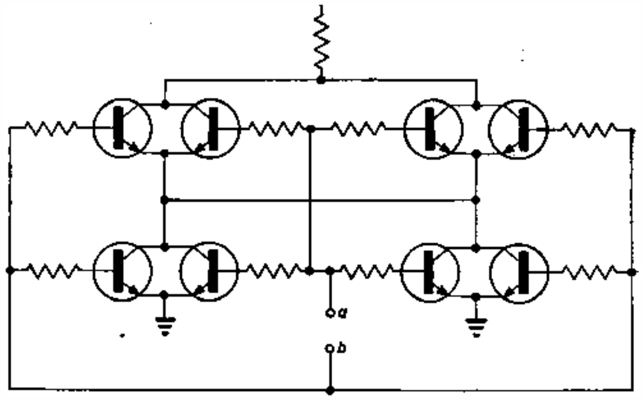


FAILURE MODE

P_1 Gate short-circuit
P_2 Gate open-circuit
P_3 a Open-circuit
P_4 b Open-circuit

GATE FAILURE MODES	PROBABILITY OF SET WORKING
P_1P_2	1
P_1P_3	1
P_1P_4	1
P_2P_3 or P_2P_4	$\frac{1}{3}$
P_3P_4	1
P_1^2	$\frac{2}{3}$
P_2^2	$\frac{1}{3}$
P_3^2 or P_4^2	$\frac{1}{3}$
P_1^3 or P_2^3 or P_3^3 or P_4^3	0
$P_1^2P_2$ or $P_1^2P_3$ or $P_1^2P_4$	$\frac{2}{3}$
$P_2^2P_1$	$\frac{1}{3}$
$P_2^2P_3$ or $P_2^2P_4$	0
$P_3^2P_1$ or $P_4^2P_1$	$\frac{1}{3}$
$P_3^2P_2$ or $P_4^2P_2$	0
$P_3^2P_4$ or $P_4^2P_3$	$\frac{2}{3}$
$P_1P_2P_3$ or $P_1P_2P_4$	$\frac{1}{3}$
$P_1P_3P_4$	1
$P_2P_3P_4$	0

Fig. 4. Redundant NOR gates, series-parallel



FAILURE MODE

P_1 Gate short-circuit
P_2 Gate open-circuit
P_3 a Open-circuit
P_4 b Open-circuit

GATE FAILURE MODES	PROBABILITY OF SET WORKING
P_1P_2 or P_1P_3 or P_1P_4	1
P_2P_3 or P_2P_4	$\frac{2}{3}$
P_3P_4	1
P_1^2	$\frac{1}{3}$
P_2^2 or P_3^2 or P_4^2	$\frac{2}{3}$
P_1^3 or P_2^3 or P_3^3 or P_4^3	0
$P_1^2P_2$ or $P_1^2P_3$ or $P_1^2P_4$	$\frac{1}{3}$
$P_2^2P_1$	$\frac{2}{3}$
$P_2^2P_3$ or $P_2^2P_4$	0
$P_3^2P_1$ or $P_4^2P_1$	$\frac{2}{3}$
$P_3^2P_2$ or $P_4^2P_2$	0
$P_3^2P_4$ or $P_4^2P_3$	$\frac{2}{3}$
$P_1P_2P_3$ or $P_1P_2P_4$	$\frac{2}{3}$
$P_1P_3P_4$	1
$P_2P_3P_4$	$^2/_9$

Fig. 5. Redundant NOR gates, parallel-series

is carried out as follows:

Let P_1 = probability that a gate fails short-circuit
P_2 = probability that a gate fails open-circuit
P_3 = probability that a gate fails open-circuit on the a side
P_4 = probability that a gate fails open-circuit on the b side
Q_s = probability that a transistor fails short-circuit
Q_o = probability that a transistor fails open-circuit

By considering the eight possible modes of failure of the gate due to one or two transistor failures it is found that:

$$P_1 = 2Q_s + 2Q_o \quad Q_s + Q_s^2$$
$$P_2 = Q_o^2$$
$$P_3 = Q_o$$
$$P_4 = Q_o$$

It can be seen from the circuits of the quadded gate that in each case any one gate can fail without causing a circuit failure but that with two or three gates failing the circuit will work in some cases. The probability of working in these cases is given in the Tables in Figs. 4 and 5.

If R_g and Q_g are the probabilities that a single gate works or fails and R_s the probability that the redundant set works, then, if $f_1 f_2$ are the probabilities that the redundant set works with 2 or 3 gate failures respectively:

$$R_s = R_g^4 + 4R_g^3Q_g + 6f_1 R_g^2Q_g^2 + 4f_2 R_gQ_g^3 \quad \ldots\ldots \quad (4)$$

where:

$$Q_g = P_1 + P_2 + P_3 + P_4 \quad \ldots\ldots\ldots\ldots \quad (5)$$

If equations (4) and (5) are combined and the values of f_1, f_2 from the tables of Figs. 4 and 5 used to multiply the cross product terms then, for the circuit of Fig. 4:

$$R_s = R_g^4 + 4R_g^3Q_g + 2R_g^2(2P_1^2 + P_2^2 + 8P_3^2 + 6P_1P_2 + 12P_1P_3 + 4P_2P_3) + 4R_g\{2P_1^2(P_2 + 2P_3) + P_2^2P_1 + 2P^2_3(4P_1 + P_3) + 4P_1P_2P_3\}$$

For the circuit of Fig. 5:

$$R_s = R_g^4 + 4R_g^3Q_g + R_g^2(2P_1^2 + 4P_2^2 + 20P_3^2 + 12P_1P_2 + 24P_1P_3 + 16P_2P_3) + 4R_g\{P_1^2(P_2 + 2P_3) + 2P_2^2P_1 + 2P_3^2(5P_1 + (2/3)P_2 + 2P_3) + 8P_1P_2P_3\}$$

The results given in Table 3 are for a chain of 48 gates.

Majority Voting

The arrangement of the gates is shown in Fig. 6.

If the reliability of each of the $3n$ gates is R_c the reliability of n gates in series is R_o where $R_c = R^n$.

If Q_c is the unreliability of a chain of n gates the probability R_n that two out of three of the chains have not failed is given by:

$$R_n = R_c^3 + 3R_c^2Q_c \text{ (from equation (1))}$$
$$= R^{3n} + 3R^{2n}(1 - R^n)$$
$$= 3R^{2n} - 2R^{3n}.$$

The majority voting unit is required to give the correct output if two out of the three inputs connect. If this unit has a reliability R_m the sub-system of Fig. 6 has a reliability $R_{ss} = R_mR_n$.

If the original non-redundant system used m gates in series the redundant system would use m/n units of n gates per chain in each sub-system and the overall reliability of the system R_s is given by:

$$R_s = R_{ss}^{m/n}$$
$$= [R_m(3R^{2n} - 2R^{3n})]^{m/n}$$

TABLE 3
Gate Redundancy

FAILURE RATE		TIME	RELIABILITY OF CIRCUIT	
Transistor	*Resistor*		Fig. 4	Fig. 5
10^{-7}	0	10^{4}	0·99976	0·99984
10^{-6}	0	10^{4}	0·9762	0·9842
10^{-7}	10^{-8}	10^{4}	0·99497	0·99506
10^{-6}	10^{-8}	10^{4}	0·97141	0·97958

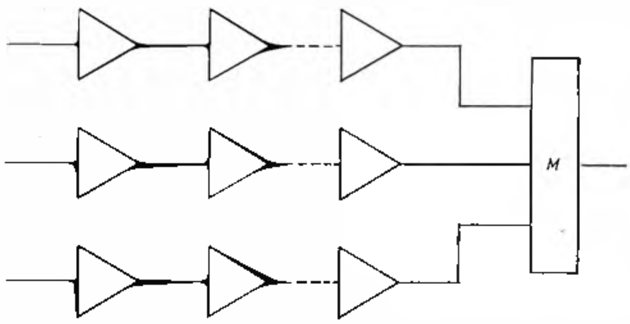


Fig. 6. Majority voting system
Three chains each of n gates in series are connected to a majority vote element

Assuming that the majority voting unit is equivalent to four gates, then:

$$R_m = R^4.$$

If:

$$R = 0{\cdot}98$$
$$m = 48.$$

Table 4 gives the reliability for different values of n.

Quadded Logic

This is a method of applying redundancy by quadruplicating each of the elementary gates of the orginal system and interconnecting them according to certain rules[2,3]. An example using NOR gates is shown in Fig. 7. For these gates the rule is that the output connexions of a set of four must follow a different pattern to the input connexions.

An analysis of the circuit shows that a single gate failure will not produce an error, and that the network will operate correctly in three cases out of six if two gates fail open-circuit or short-circuit and in all cases if one gate fails short-circuit and one open-circuit, provided that the subsequent gates have not failed. It can be shown that some extra failures in the later stages are also corrected.

The probability that a system of 4 × 48 gates will fail can be calculated approximately as follows. Since the individual gate reliability is high the probability of more than say 10 failures is very low and the following reasoning is sufficiently accurate.

If one gate has failed there are 3/191 chances that the second failure occurs in the same set of four. If the second gate is not in the same set of four as the first there are 6/190 chances that the third failure will occur in the same set as the first or second failure. Hence if n gates have failed the probability P_1 that two failures have occurred in one set of four is given approximately by:

$$P_1 = (3/191)+(6/190)+(9/189)\ldots\ldots\frac{3(n-1)}{193-n} \text{ if } n \geqslant 2$$

$$= \frac{n(n-1)}{128}$$

Since n is small the probability that more than one set of four gates has two faults is neglected.

The probability P_3 of failure due to faults in consecutive sets of gates is estimated on the assumption that the system fails if a fault occurs in each of two consecutive sets of four and one of the next two sets of four is also faulty. The probability that two out of n failures will occur in consecutive sets of gates is approximately $4(n-1)n/(192)$ and the probability that other failures will occur in the next two sets is $8(n-2)/(192)$ approximately.

Hence:

$$P_2 \simeq \frac{32\,(n-1)\,(n-2)n}{192^2}$$

Hence the probability P_3 of failure with n failures is given by:

$$P_3 = P_4P_1 + 2$$

where P_4 is the probability of failure with two failures in one set of four.

If P_n is the probability of exactly n failures the probability of system failure Q_{ss} is given by:

$$Q_{ss} = \sum_{0}^{n} P_n\,(P_1 + P_2) \text{ (if } n < 10)$$

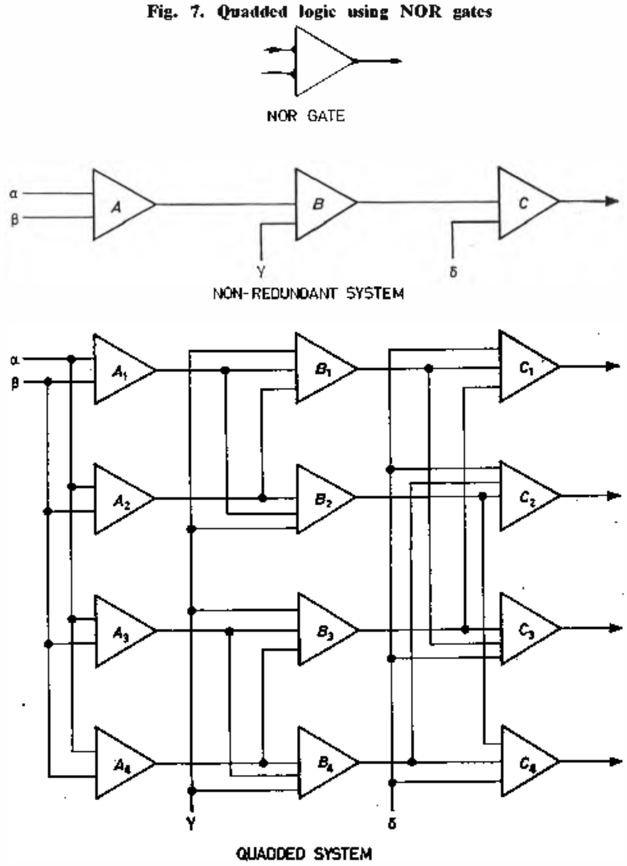
The probabilities of 0, 1, 2 .... failures are given by the terms in the following expansion[1]:

$$e^{-n\lambda t}\left(1 + n\lambda t + \frac{n^2\lambda^2 t^2}{2}\right) + \ldots.$$

TABLE 4

t	n	48	16	12	8	6	NON REDUNDANT
0·01	R_s	0·298	—	0·357	0·365	0·35	0·38
0·001	R_s	—	0·935	0·961	0·948	—	0·908

Fig. 7. Quadded logic using NOR gates



where λ = gate failure rate

n = number of gates

t = time.

It can be seen from the circuit that each three-input gate fails if any one transistor fails short-circuit but only fails in one-third of the cases where one transistor fails open-circuit. Each gate always fails if two transistors fail open-circuit.

If Q_o and Q_s are the open- and short- circuit failure rates of the transistor the gate failure rates $Q_g \simeq Q_o + 3Q_s$.

The approximate reliability of the system R_{ss} is 0·99978 if $\lambda = 10^{-7}$, $t = 10^4$ and 0·95966 if $\lambda = 10^{-6}$, $t = 10^4$, where λ is the transistor failure rate and t is time in hours. As before the assumption is made that the transistor is three times as likely to fail open-circuit as short-circuit.

Assembly and Testing Techniques

It is possible to build any of the circuits described using conventional components or integrated circuits, both thin film and solid circuits. Where space and weight are the prime requirements solid circuits are to be preferred, but at present the power consumption is higher than for either of the other techniques.

Since, by definition, a system employing redundancy will work if one or more components have failed, a functional test on the complete equipment will not show whether all the components are working. Systems employing component or gate redundancy can only be checked if each component or gate can be checked. Hence component redundancy cannot be used with integrated circuits if checking can only be performed at system or subsystem level. The practical difficulties of checking individual gates with assemblies of integrated circuits are such that this technique will not usually be practicable if a check on redundancy is required after assembly.

Majority voting systems can be checked by adding an error detector to each majority vote unit. Some means of checking the error detector itself must be provided. It will be noted that this is about as complicated, hence as unreliable, as the majority voting unit itself.

It is also possible to check a majority logic system if separate power supplies are used for each chain and for the majority voting units.

Quadded logic is easily checked. If four power supplies are used, each feeding one gate out of each set of four, the system will work if any three of the power supplies are off, provided all of the gates are functioning correctly. More complicated systems can involve different combinations of power supplies but it seems certain that checking by switching supplies is always possible.

TABLE 5

SYSTEM	APPROXIMATE RELATIVE POWER
Non redundant	1
Component redundancy	1 (transistors only) 4 (all components)
Gate redundancy	1 (transistors only 1·2 (all components)
Majority vote ($\frac{2}{3}$)	3
Quadded logic	4

TABLE 6
Probability of Failure

SYSTEM λt	10^{-3}	10^{-2}
Non-redundant	0·092	0·621
Majority vote ($\frac{2}{3}$)	0·039	0·635
Quadded logic	0·00022	0·0403
Gate redundancy	0·00016	0·0155
Component redundancy	0·00013	0·013

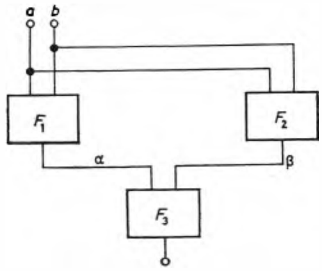


Fig. 8. Triangular network

TABLE 7

FAILURE MODE	F_1	F_2	F_3
None	$\overline{a+b}$	$\overline{a+b}$	α,β
1 Transistor o.c.	$\bar{a}$ or $\bar{b}$	$\bar{a}$ or $\bar{b}$	0
2 Transistor o.c.	1	1	0
1 Transistor s.c.	0	0	α or β
2 Transistor s.c.	0	0	1
1 Transistor o.c. and 1 Transistor s.c.	0	0	0

Relative Power Requirements

If it is assumed that computer speeds are so low that the extra power demanded to compensate for the increased stray capacitances of the redundant circuits is negligible, then Table 5 gives the relative supply powers required for the different systems.

The assumption is also made that the required fan-out can be obtained without an increase in power. If this is not the case the increase will be greatest for component and gate redundancy and least for quadded logic.

Conclusions

The best type of redundancy to use in any particular application depends on the type of component to be used, the maximum power available, and the constructional techniques adopted.

If individual conventional components are to be used and checking facilities for each component are possible then redundancy at component level gives the maximum improvement in reliability.

If completely sealed equipments are to be manufactured and a check on the system performance after final assembly is required then quadded check or majority voting offer the only simple solutions. The gain in reliability is very large with quadded logic but the power required has increased four times.

Table 6 summarizes the reliabilities of the systems described.

Acknowledgments

The author is indebted to Mrs. T. Klaschka for programming and computing most of the numerical results and also for some useful discussions on the theoretical aspects of redundancy.

Appendix

(A) Reliability of Series-Parallel Components

The circuit is shown in Fig. 1(a).

The reliability can be calculated from equation (1) if Q is replaced by $Q_o + Q_s$ and each term is multiplied by the probability of working as given in the table in Fig. 1(a).

Hence if R_c = reliability of set

R = reliability of one component

Q = unreliability of one component

Q_s = probability of one component failing short-circuit

Q_o = probability of one component failing open-circuit

$$R_o = R^4+4R^3Q+6R^2\left(\tfrac{1}{3}Q_o^2+2Q_oQ_s+\tfrac{2}{3}Q_s^2\right) + 4R\left(3\tfrac{1}{3}Q_o^2Q_s + 3\tfrac{2}{3}Q_oQ_s^2\right)$$
$$= R^4+4R^3Q+2R^2Q_o^2+12R^2Q_oQ_s + 4R^2Q_s^2+4RQ_oQ_s+8RQ_oQ_s^2$$

(B) TRIPLE REDUNDANCY

A single gate may be replaced by three gates connected[3] as in Fig. 8 or by three in series or parallel. Series or parallel arrangements are not usually practical.

In the circuit of Fig. 8, F_1, F_2 and F_3 are arbitrary functions, the combination performs the function required. The individual functions should be chosen so that the triple works if one or more failures occur in the individual gates.

The best arrangement for a NOR function was found to be when F_1F_2 were NOR gates and F_3 an AND gate.

Table 7 can now be constructed, from an analysis of the practical circuit derived from Fig. 8.

From this it can be seen that the system will work if one transistor in F_1 or F_2 is open-circuit, or if two transistors in F_1 or two in F_2 are open-circuit, if one in F_1 and a particular one in F_2 are open-circuit. The system will also work if one transistor in F_3 is short- circuited or one in F_3 short-circuited with any combination of open- and short-circuits in F_1 or F_2.

If Q is probability of a transistor failure, Q_o the probability of an open-circuit failure and Q_s the probability of a short-circuit failure, then:

$$Q = Q_o + Q_s$$
$$R = 1 - Q.$$

Expanding $(R + Q)^6$:

$$(R+Q)^6 = R^6+6R^5(Q_o+Q_s)+15R^4(Q_o^2+2Q_oQ_s+Q_s^2)$$
$$20R^3(Q_o+3Q_oQ_s+3Q_s^2Q_oQ_s^3) + \ldots \quad \text{(B.1)}$$

If each term is multiplied by the probability, that the system will work if the failure combination given by that term occurs (from Table 7), we have for the reliability R_s of the system:

$$R_s = R^6+4R^5Q_o+2R^5Q_s+R^4(4Q_o^2+4Q_oQ_s+4Q_s^2) + R^3(2Q_o^2Q_s+2Q_oQ_s^2+2Q_s^3) \quad \ldots \text{(B.2)}$$

The reliability of a single gate R_g is given by:

$$R_g = R^2 \quad \ldots\ldots \text{(B.3)}$$

From equations (B.2) and (B.3):

$$R_s/R_g = 1-2Q_s-2Q_o^2-2Q_oQ_s+4Q_s^2+4Q_o^2Q_s - 2Q_oQ_s^2-4Q_s^3+Q_o^4+4Q_oQ_s^3+Q_s^4 \quad \ldots \text{(B.4)}$$

Since $Q_s \ll 1$ and $Q_o \ll 1$

$$R_s/R_g \simeq 1 - 2Q_s$$

Hence the triple gate network is worse than the single gate, since R_s/R_g is always less than unity if Q_o and Q_s are small.

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An Automation Instrument Data Service

As a result of the recent evaluation of industry's needs for information, and the study of methods to satisfy these needs, the GKN subsidiary Indata Limited has developed and is now marketing a commercial industrial information service.

The Automation Instrument Data Service, which will come into operation in May of this year, is the product of close co-operation between the Scientific Instrument Research Association and Indata Limited. Aimed at an industrial problem that has been well defined and plumbed through exhaustive market surveys, this new service will help industry in two ways—firstly as a service that provides selective 'current awareness' of new products to the engineers and technologists using the many components and materials introduced every year, and secondly as a means of presenting manufacturers' new products to the market most in need of them. Ultimately the 'current awareness' service may be developed to incorporate a full information retrieval system based on the bank of product data that is now being built up.

The Automation Instrument Data Service is a computer-based industrial information system that has been developed by Indata Limited with the active co-operation of the Scientific Instrument Research Association. Concentrating initially on the automation, electronics and scientific instrument industries, the service offers to engineers, technologists and purchasing departments a 'current awareness' of all new products introduced in subject categories specified by subscriber.

Subscribers choose the categories of products in which they are interested from a category list of 309 engineering subjects—this list being backed by an index of some 850 terms. Once chosen and entered into the Indata computer, these categories, which are collectively known as the subscriber's profile, are used to match the subscriber's needs against incoming new products during each weekly computer run, and thus determine the product cards he will receive.

Each subscriber to the service receives a modular library of loose-leaf folders in which standard sized 80-column computer cards are housed. These cards, one for each new product, are sent—on a weekly basis—to the subscriber as and when products are received within the range of subjects specified by him. They contain a full engineering specification with an illustration and a free-text description of up to 800 words.

The subscriber also receives master lists of the cards he is to receive each week, and profile lists for up to five other individuals within the organization who wish to kept informed of products falling into sub-sets of the range the subscriber has specified. Indata maintains a servicing department to ensure that all recently received cards are filed.

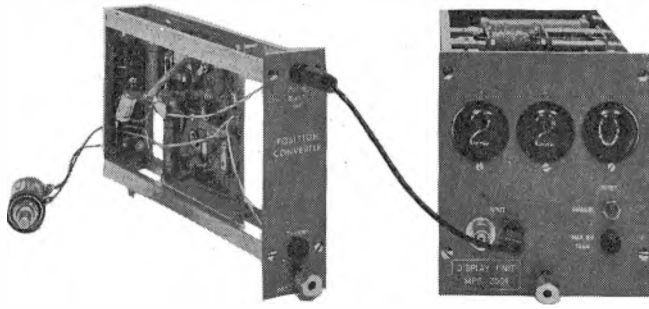
Feedback, to manufacturers, of interest in products is facilitated by a tear-off stub that forms part of each product card. This stub contains boxes which the subscriber marks acording to whether he requires quotations, a representative to call or a demonstration; boxes are also provided which allow him to identify himself as a named subscriber. When received at Indata the marks on the stubs are read by the computer which prints out manufacturer advice lists.

The library of folders, which are matched to the subscribers' needs, also contain cards that enable him to notify Indata of any change of profile. When one of these is filled in Indata automatically changes the profile and provides a three-month back-log of cards in any new categories.

Comprehensiveness of the range of product data supplied by the service is maintained by obtaining the co-operation of the maximum number of manufacturers who do not pay to have their products included in the system. Accuracy is maintained by a team of graduate engineers who edit and vet the incoming data to ensure that like products are described in unambiguous engineering terms.

Initially offering a low cost 'current awareness' service, the system developed by Indata has been planned from the start to be capable of expansion into an advanced information retrieval service with 'real-time' capabilities. Using the data-bank, which will be produced by the first phase, an evaluation program can be initiated to investigate the use of directly linked remote terminals providing 'real-time' access to the data-bank.

A Convertor for Automatic Recording of Mechanical Position



By U. Kracht*

A simple monostable multivibrator circuit is used to convert a resistance value to a square wave whose length is proportional to the resistance. On a potentiometer the slider is attached to the mechanical distance movement through a suitable gear, so that the assembly acts as a mechanical position to time convertor. The square pulse, whose start is initiated by an external trigger pulse determining the moment at which the mechanical position is measured, is converted to a train of pulses suitable for display on a scaler or connexion to a data acquisition system. The number of pulses is then proportional to the mechanical position.

The property of the same multivibrator circuit is briefly considered as a voltage-to-time convertor.

(Voir page 276 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

AMONG the possible forms of parameter which have to be recorded in data acquisition systems is the position of mechanical parts, such as screens, slits, coaxial phase shifters and r.f. cavity tuner plates, to mention just a few. A common feature of many such mechanical movements is that they are fairly slow and that the total movement from one end to another is fixed.

By connexion of a potentiometer through a suitable gear, a reliable and sufficiently precise indication of the mechanical position can be obtained. Linear potentiometers also are available for direct connexion to a moving object. Such systems have the important feature of indicating correctly even if the connecting cables have been disconnected for some time and reconnected. This is not the case for a pulse generator connected to the mechanical movement, and often such a device does not give the same number of pulses counting in the reverse direction. Furthermore, the potentiometer is cheaper than a pulse generator.

This article describes a convertor which converts the resistance value to a period of time, or rather to a train of pulses whose number is proportional to the resistance value. This train has a frequency of 100kHz and consists of from 20 to a maximum of 220 pulses, to be compatible with the data logging system used at present in the CERN linear accelerator (LINAC[1]). To define which cycle in the CERN Proton Sychrotron (CPS) the mechanical position relates to, the convertor needs a trigger signal from the CPS timing system. If a display scaler only is used, giving a direct visible indication of the mechanical position, the trigger signal can conveniently be used to reset the scaler after each cycle. For the display of many potentiometer values, only one convertor and one display scaler are needed.

Referring to Fig. 1, which is a simplified form of the actual convertor shown in Fig. 2, the principle of the convertor can be described as follows. The circuit is a normal monostable multivibrator which after being triggered, will give an output pulse whose length is determined by R (C being kept constant), and here R is the potentiometer connected to the mechanical movement. VT_3 is a constant current source whose current is determined only by R. This is necessary in order to linearize the relation between the value of R and the pulse length, and also to eliminate any influence on the circuit of long cables between the convertor and R. Several microfarads of capacitance can be connected between either of the two potentiometer cables and earth, or across the potentiometer without any influence on the linearity or precision. A simple, low-pass filter is connected at the input to eliminate influences caused by heavy pulse currents which are often present in accelerators. Fig. 3 shows the actual waveforms on transistor VT_4 for different values of R. VT_1 (Fig. 2) is the input transistor and R_3 introduces a threshold for

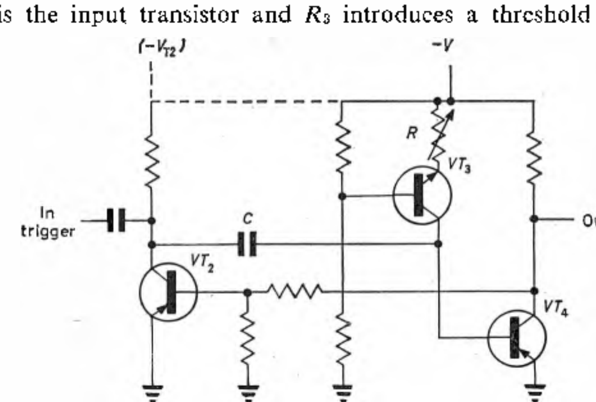


Fig. 1. Basic monostable multivibrator circuit

the trigger signal. With this arrangement, noise immunity is increased and the trigger signal shape has very little effect on the convertor. VT_5 minimizes load variation effects on the convertor. From Fig. 2, it is seen that the output from the multivibrator is used to open up a gate for 400kHz pulses from a quartz crystal oscillator. The uncertainty in the gate (error always one to two pulses) is greatly reduced by dividing by four afterwards (VT_8, VT_9 and VT_{10}). VT_{11} is the output stage, giving about 10V unloaded and about 5V with 75Ω load. Silicon passivated epitaxial transistors are used throughout to minimize temperature effects. In the actual convertor part, a low supply voltage and low currents eliminate warming-up effects. The diode MR_3 compensates for temperature drift in VT_3.

As it is not possible to make the value of R (the

The above photograph shows, from left to right the ten-turn potentiometer which normally is connected to the mechanical movement, the convertor and a display unit (scaler). The necessary trigger pulse connexions are not shown

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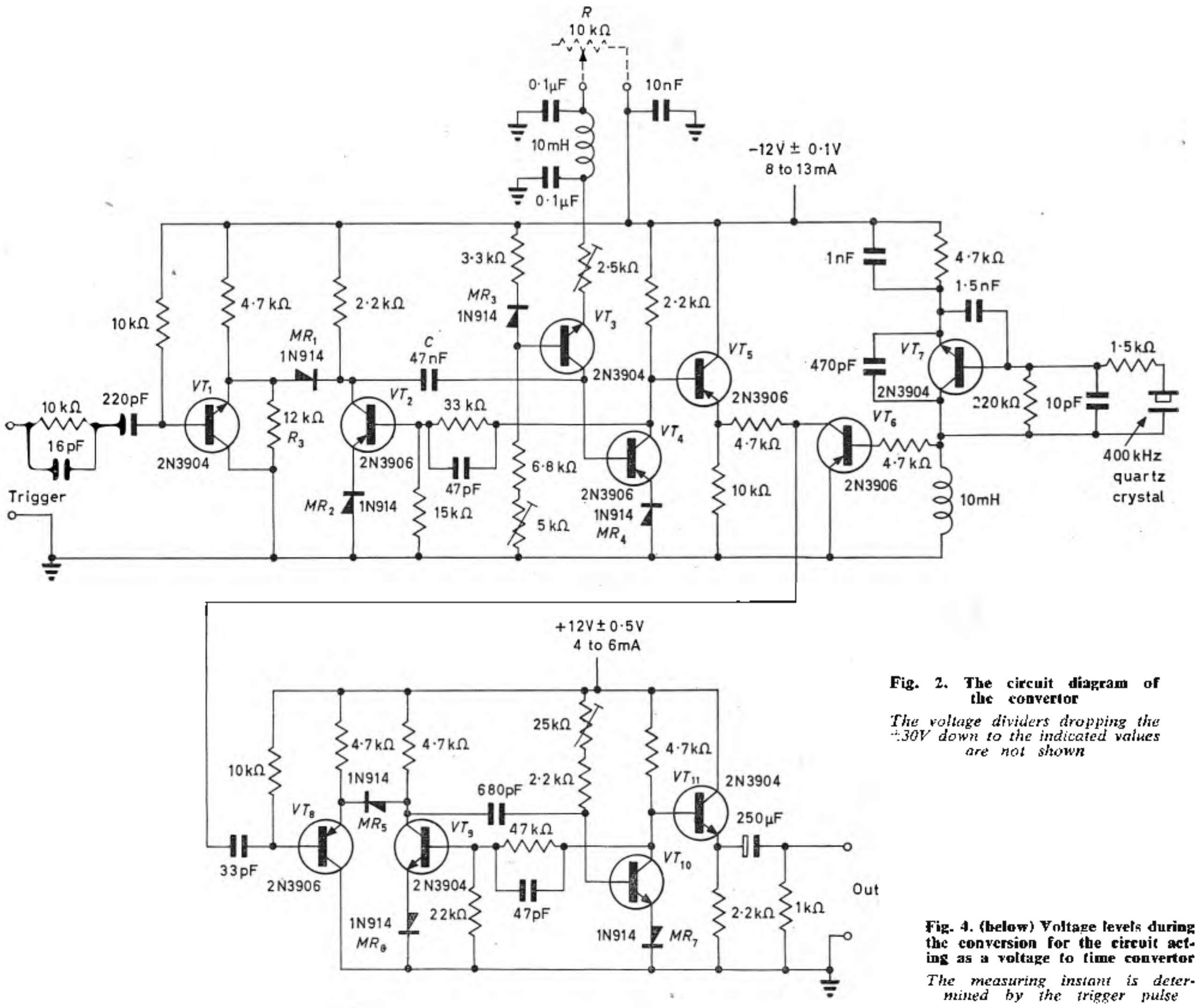

Fig. 2. The circuit diagram of the convertor

The voltage dividers dropping the ±30V down to the indicated values are not shown

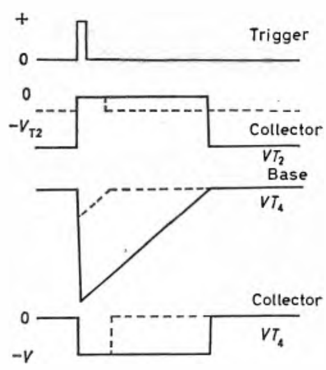

Fig. 4. (below) Voltage levels during the conversion for the circuit acting as a voltage to time convertor

The measuring instant is determined by the trigger pulse

potentiometer) zero without exceeding the rating of VT_3, a preset potentiometer RV_1 is connected in series with R. This fixes a certain minimum pulse length for zero value of R, but this is not very serious. Actually, this effect is of advantage, because one has an indication as to whether the potentiometer is connected or not, and in the computer programme this can easily be dealt with.

The convertor, which is shown on page 250, together with a display scaler, has the following performance:

Input resistor value	0 . . 10 000Ω
Output pulses for 0Ω	20
Output pulses for 10 000Ω	220
Output frequency	100kHz
Output voltage	10V
Trigger signal, minimum	0·5μsec, 10V
Temperature dependence, less than 1 pulse/10°C	
Ambient temperature range	10 to +70°C
Supply voltages/current	+30V/10mA, −30V/18mA

Fig. 4 shows the voltage on the collectors of VT_2 and VT_4 as a function of time and varying supply voltage to VT_2. From this it will be seen that the multivibrator can be used not only as a resistance-to-time convertor, but also as a voltage-to-time convertor.

A free running resistance-to-time and voltage-to-time convertor is being used in CERN for a distributed carrier multiplex system[2]. The resistance value (potentiometer) is the same for the two convertors.

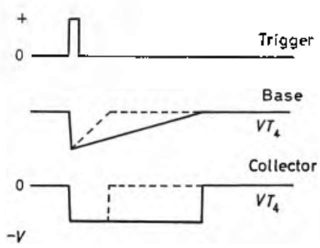

Fig. 3. (above) Voltage levels during the conversion for the circuit acting as a resistance to time convertor

The measuring instant is determined by the trigger pulses

Acknowledgments

The author wishes to thank CERN for the kind permission to publish this article and Mr. C. S. Taylor, head of the Linac Group, for his stimulating interest in this work. Gratitude is also expressed to Messrs. B. Sagnell, P. Bramham and K. Unser.

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Constant Voltage Double Pulse Generators for Use in Electrophysiology

By A. F. Lewis*, B.Sc., Ph.D., M.B., B.Ch.

Design features are discussed for double pulse generators for use in electrophysiology with emphasis on the versatility required to meet the wide range of possible applications. A reasonably simple solid state circuit is described to provide constant voltage paired rectilinear pulses of independent amplitude and duration. The output impedance is very low (5Ω) and the timing circuits are adequately stabilized against temperature and supply voltage variations so that the unit is suitable for battery operation. Rise and fall times are 5 and 10μsec respectively and the whole unit is sufficiently robust electrically to withstand prolonged abuse in undergraduate classwork.

(Voir page 276 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 277)

THE cell membrane of resting excitable tissue (nerve or muscle) has a potential of +80mV outside relative to the inside. This potential is determined by the equilibrium state of the ionic concentrations either side of the membrane, i.e. inside and outside the cell, together with the permeability of the membrane to these ions.

The application of mechanical, chemical or electrical energy to the tissue may result in a considerable increase in the characteristic permeability of the membrane for sodium ions. This change in the equilibrium condition causes a transient reversal of the resting membrane potential which is known as an action potential (Fig. 1). The function of nervous tissue is to propagate this action potential along its length. In muscle the action potential initiates the process of contraction.

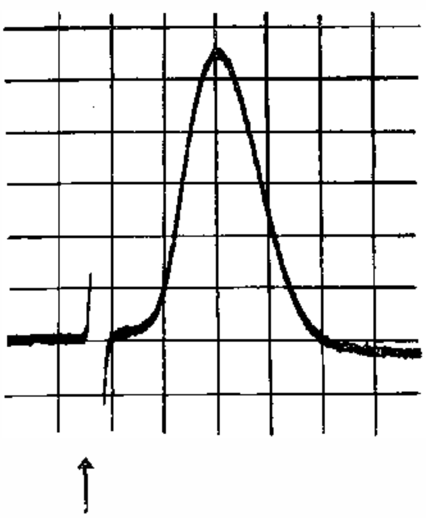

Fig. 1. A nerve action potential following an electrical stimulus

The differentiated stimulus appears where marked by the arrow and the reversal of the membrane potential follows over the next 2msec. The total potential change is 140mV but because of the high source impedance only 2·8mV appears across the recording electrodes.

X axis: 0·5 msec/cm. Y axis: 0·5mV/cm.

Electrophysiological techniques measure the excitability of these tissues in terms of the amount of applied electrical energy necessary to cause this change (the threshold stimulus). Like electrical switching circuits, excitable tissues have a recovery period during which they are totally or relatively refractory to further stimuli. This refractory interval is measured by application of paired pulses (Fig. 2).

These tissues are commonly immersed in oxygenated electrolyte solutions during investigation and it is the low impedance of this solution that determines the output requirements of the pulse generator. A negligibly low generator output impedance avoids the necessity to monitor the actual applied voltages across the preparation.

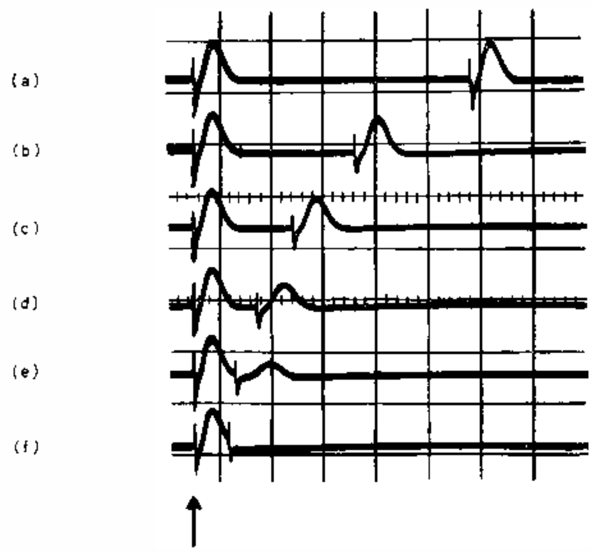

Fig. 2. The measurement of refractory period of nerve

Two spaced identical stimuli are applied to the nerve, the first where marked by the arrow and the second between 1·5 and 10·5msec later. The stimulus is recorded as a small diphasic artefact, normally followed by an action potential. In traces D, E and F the response is impaired or absent due to the refractory state of the nerve.

X axis: 2msec/cm. Y axis: 2mV/cm.

Battery Operation

The advantages of battery operation are many. The complete portability of battery equipment is enhanced by the omission of the heaviest and bulkiest components—the mains transformer and associated rectifying and smoothing components. The presence of mains operated equipment near the tissue preparation considerably in-

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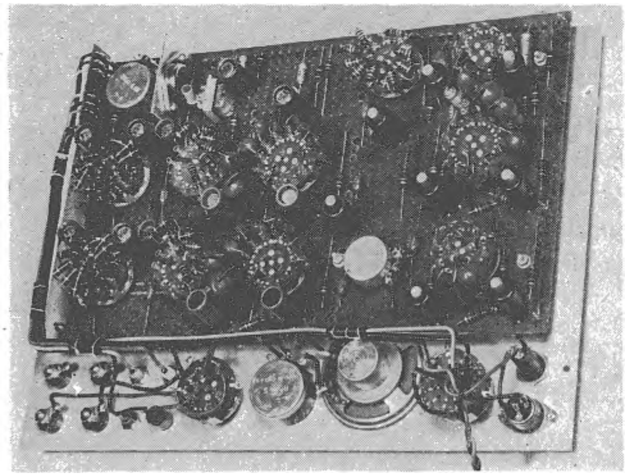


Fig. 3(a). The complete stimulator assembled on Veroboard which is then mounted on the instrument panel

Only the battery and decoupling capacitor are not shown.

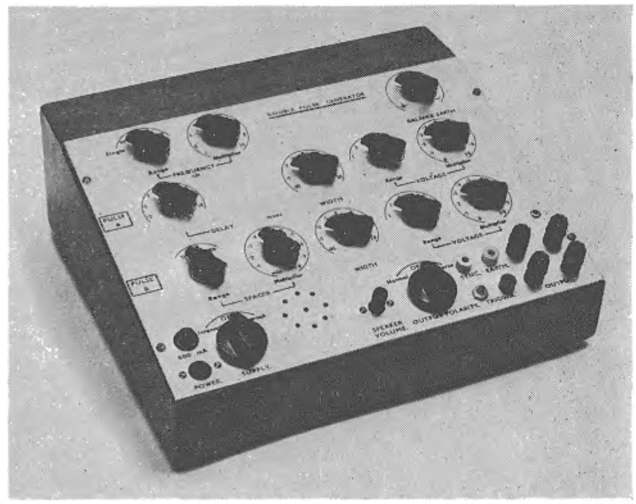


Fig. 3(b). The complete instrument

creases the problem of discriminating against common-mode 50Hz a.c. interference on the biological signal. Furthermore battery operation allows the whole apparatus to float in relation to earth so that the pulse output is independent of earth (except for the inevitable small capacitive effect). This isolation of the stimulating pulse from earth allows balancing to minimize the stimulus artefact at the recording electrodes. Thus the use of the relatively unsatisfactory pulse isolating transformer is unnecessary and the more complex methods of r.f. transformer isolation are avoided.

The voltage of the battery is decided by the maximum pulse amplitude required whereas the timing units can economically be supplied from a stabilized supply of about 12V. Multiple battery supplies are best avoided in apparatus of this sort and so to minimize power waste a battery of 45V was selected. Many commercially available batteries of this voltage and above are becoming obsolete with the decreasing use of valve operated equipment but production of the Ever Ready B104 45V unit is assured for the next five years at least.

Specification

The following specification agrees with the Medical Research Council recommendations[1] for electrodiagnostic stimulators, except for the values of pulse width and amplitude. Their recommended amplitude of 200V was in relation to an output impedance of up to 500Ω. Since semiconductor circuits can realize very much lower output impedances a corresponding reduction in amplitude is possible.

The very large pulse durations recommended (up to 300msec) have little application in experimental neurophysiology and they have not been included. An alternative pulse generator circuit giving pulse durations up to 1sec without using electrolytic capacitors is also described.

(a) Frequency range: 0·1 to 1 000Hz.

(b) Pre-pulse for oscilloscope synchronization.

(c) Loudspeaker monitor.

(d) Trigger for single pulses or trains of pulses.

(e) Delay range after pre-pulse: 0·2 to 50msec.

Fig. 4(a). Arrangement of circuit sub-units

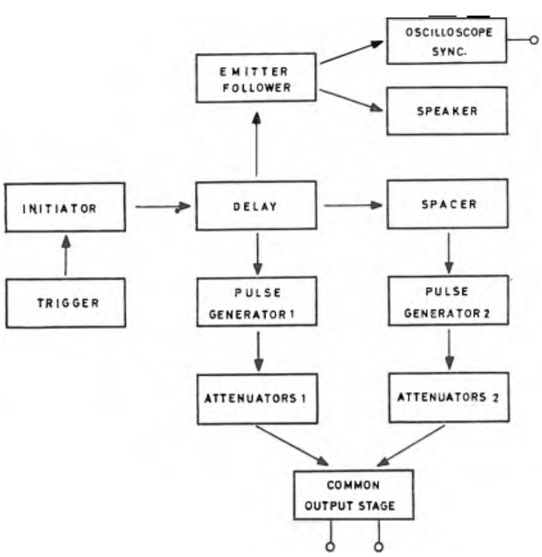
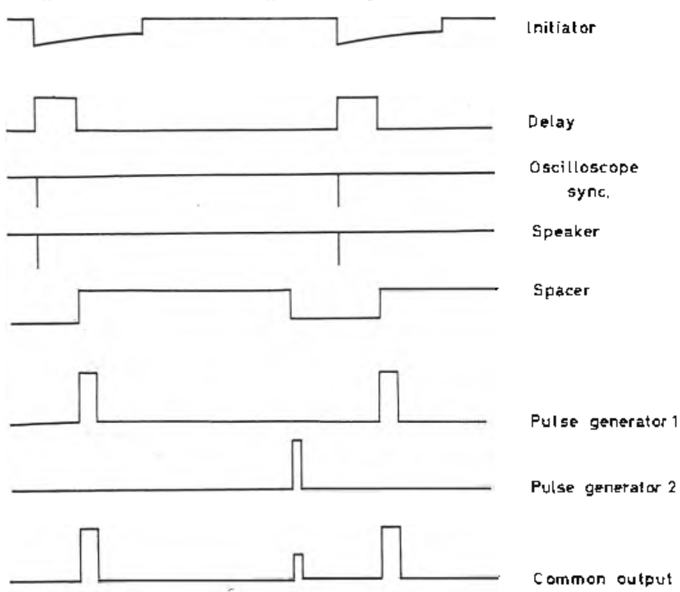


Fig. 4(b). The waveforms generated by the sub-units of the stimulator



(f) First pulse generator: duration 50μsec to 20msec amplitude 0 to 40V.

(g) Spacer ranges: 0·3 to 10msec
3 to 100msec
30 to 1 000msec

(h) Second pulse generator with identical specification to first.

(i) Common low impedance output stage.

It is essential that the common output stage prevents summation of overlapping pulses.

As long as the production of electrophysiological equipment at reasonable cost remains commercially unattractive, laboratories will continue to assemble their own. There is unfortunately a large hiatus between the published designs available and the practical problems of construction. The aim of this article is to provide sufficient detail of the sub-units to allow versatile assembly of a stimulator.

Construction

Most compact circuits are readily constructed upon Veroboard. For production of a small number of laboratory instruments capable of subsequent modification, the convenience of this material is probably greater than that of a printed circuit. The limiting design factor is the layout of the control panel. Controls should be about 2in apart and may be mounted through the Veroboard thus eliminating the need for wiring looms with their troublesome capacitive effects. The circuit to be described is constructed upon 0·2in pitch Veroboard measuring $10\frac{3}{4}$in $\times$ $6\frac{1}{2}$in. All controls and components are mounted on the Veroboard or front panel with the exception of the battery and decoupling capacitor (Fig. 3(a) and (b)). The relationship of the various units is shown in Fig. 4 and the complete circuit diagram in Fig. 5.

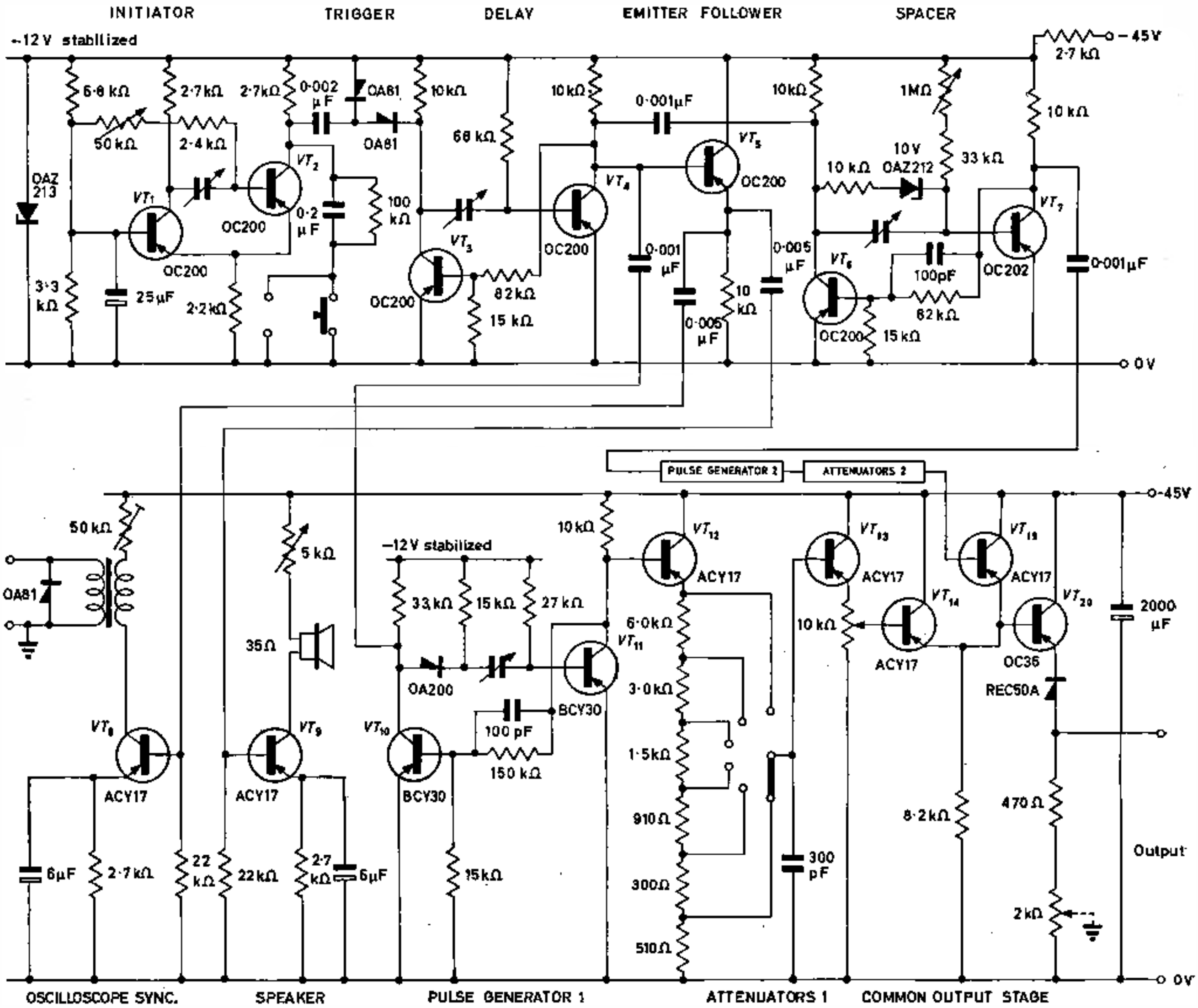
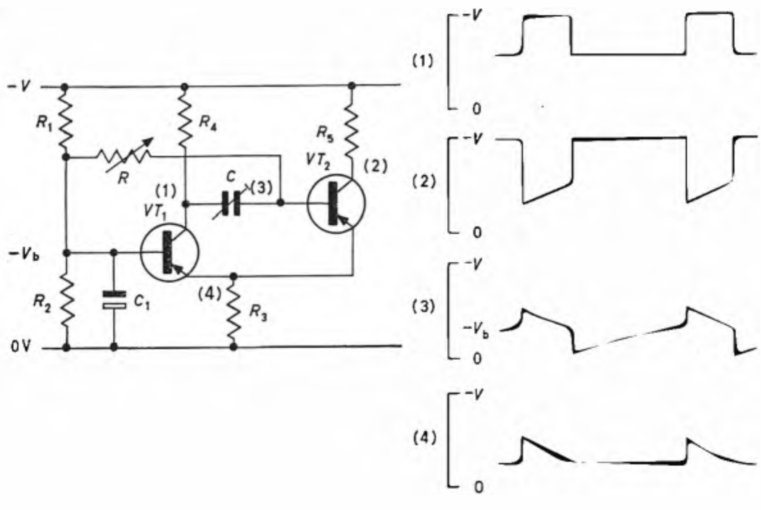


Fig. 5. The complete circuit

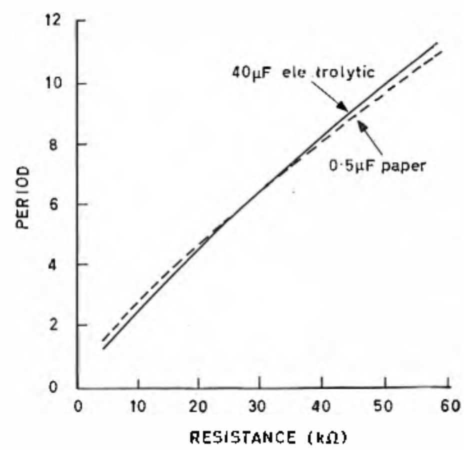
Semiconductors VT_{15} to VT_{18} form part of the second pulse generator and attenuator which are not shown in detail. The potentiometer across the output is to allow balancing about earth and is optional.

Initiator

The initiator is designed to produce a positive rectilinear pulse whose leading edge is differentiated to drive the subsequent timing units. The pulse need not be perfectly rectilinear nor is a mark-to-space ratio of unity necessary. Collector-base coupled astable multivibrators



(a)



(b)

Fig. 6(a). The circuit of the initiator
The output is taken from the collector of VT_2 and diode gated so that the positive-going leading edge drives the delay circuit.

Fig. 6(b). The relationship between oscillating period and timing resistor in the initiator for a paper and electrolytic capacitor
The period is in arbitrary units.

require ganged controls and are unnecessarily cumbersome for this purpose. The emitter coupled astable shown in Fig. 6(a) has a single capacitor C which is switched for the various ranges and a potentiometer R for adjustment within these ranges.

The potential of the base of VT_1 is clamped at $-V_b$. When VT_1 is saturated the capacitor C holds VT_2 in the cut off state while it discharges through R, there being negligible leakage through VT_2. When the base of VT_2 reaches $-V_b$ this transistor comes on and emitter re-

TABLE 1
RC values for timing units

INITIATOR

Frequency range (Hz)	*Capacitor* (μF)	*In lieu of 50kΩ potentiometer* *Frequency multiplier*	*Series resistors* (Ω)
0·1 — 1	40*	× 10	0
0·1 — 10	4*	× 9	430
1 — 0·5	0·5	× 8	680
10 — 1000	0·05	× 7	820
		× 6	1·1 k
		× 5	1·8 k
		× 4	2·7 k
		× 3	4·7 k
		× 2	11 k
		× 1	33 k

*Selected electrolytic or tantalum type

DELAY

Delay (msec)	*Capacitor* (μF)
0·2	0·005
0·5	0·01
1	0·02
2	0·04
5	0·1
10	0·2
20	0·4
50	1·0

SPACER

Range (msec)	*Capacitor* (μF)
0·3 — 10	0·015
3 — 100	0·15
30 — 1000	1·5

PULSE GENERATOR

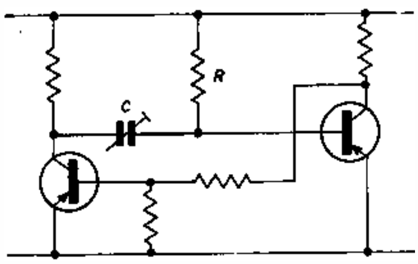
Pulse width (msec)	*Capacitor* (μF)
0·05	0·002
0·1	0·005
0·2	0·01
0·5	0·025
1	0·05
2	0·1
5	0·25
10	0·5
20	1·0

ALTERNATIVE PULSE GENERATOR

Range (msec)	*Capacitor* (μF)
0·1 — 10	0·015
1 — 100	0·15
10 — 1000	1·5

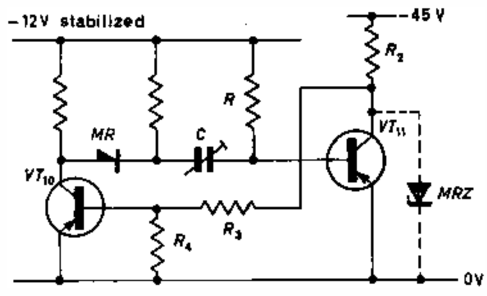
generation cuts off VT_1 so that its collector and thus also the base of VT_2 move negative. The capacitor C now charges through R and VT_2.

Thus the off period of VT_2 is proportional to R (using silicon transistors with negligible current leakage) while the on period is determined by the combined current through R and the base of VT_2. If VT_2 has a large d.c. gain the oscillating period/resistance plot is very nearly linear.



(a)

Fig. 7(a). A simple monostable ciruit



(b)

Fig. 7(b). The improved monostable circuit (Dodgson) used as the pulse generator

Pulses up to 20msec duration are available without using electrolytic or bulky capacitors. The Zener diode across the output may be used to stabilize pulse amplitude against supply voltage variation.

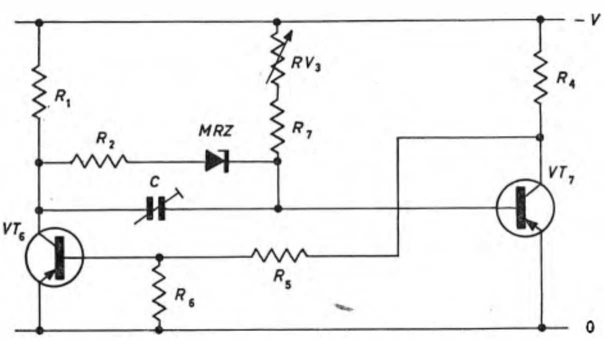


Fig. 8. A modified monostable circuit (Bannister) used as a spacer between pulses

This provides a wide range of continuous control. The Zener voltage of MRZ should be about 80 per cent of the supply voltage −V and the value of R_2 calculated to provide correct biasing of VT_7.

The oscillating period is moderately dependent upon supply voltage (which should be stabilized) and transistor parameters. Initial setting up is accomplished by adjusting R_2 or R_3. The gain of VT_2 must be sufficient to allow oscillation at the maximum value of R, i.e. the lowest frequency of the range.

Electrolytic capacitors are necessary for the lower frequencies and because of their leakage a departure from linearity of the period/resistance plot exists. This is shown in Fig. 6(b).

The relationship of frequency to resistance is hyperbolic and the closest commercially available potentiometer has an inverse logarithmic law. A stepped resistance control may be preferable and the values are given in Table 1.

The output from the collector of VT_2 is diode gated to the delay circuit.

Trigger Circuit

Grounding of the collector of VT_2 of the initiator to a.c. allows oscillation to continue but prevents drive to the delay and subsequent circuits. The grounding capacitor should be many times larger than the coupling capacitor and a 100kΩ resistor across the capacitor ensures its rapid discharge between pulses. A time-constant of 20msec is sufficient to suppress drive up to 1 000Hz. A normally closed switch across the external trigger contacts may be used for keying trains of stimuli.

With the initiator rendered inoperative by switching out the timing capacitors, the trigger button is used for initiating single stimuli, these occur on the make. Care is necessary in selecting suitable trigger switches that are free from bounce. A morse key is very satisfactory.

Delay

This is a straightforward monostable giving switched delays from 0·2 to 50msec without using electrolytic capacitors. The output is taken from the collector of VT_4 and is a negative rectilinear pulse of nearly 12V amplitude.

Precise values of delay are not necessary nor any measure to confer stability other than the use of a stabilized supply line and silicon transistors.

The leading negative edge of the output is used to drive two class-B amplifiers, VT_8 and VT_9, one supplying a loudspeaker monitor and the other a transformer isolated pre-pulse for oscilloscope synchronization. As both these amplifiers feed inductive elements they are isolated from the delay by an emitter-follower.

The positive-going trailing edge of the delay pulse is differentiated to drive the first pulse generator and the spacer unit. Note that the auditory monitor is synchronous with the oscilloscope sweep and precedes the stimulus output by the width of the delay pulse.

Oscilloscope Synchronizing Pre-pulse

The output of the transformer in the class-B amplifier VT_8 is rendered monophasic by a diode and the amplitude limited by the preset potentiometer in series with the transformer primary winding. The pulse is about 0·1msec duration and up to 40V in amplitude. As one side of the transformer secondary is connected to earth on the oscilloscope the interwinding capacitance of this transformer will limit the final capacitance to earth of the stimulator output. A suitable screened transformer should be used. The amplitude of this pulse should be limited to that adequate for oscilloscope synchronization as large pulses are liable to cause an induced artefact in the output stage.

Pulse Generator

The pulse generator is required to produce a rectilinear pulse of constant width and amplitude. In the simple monostable circuit of Fig. 7(a) the main causes of variation in these parameters are:

(a) Temperature and aging effects on the timing capacitor which are largely overcome by avoiding the use of electrolytic types.

(b) Collector saturation voltage variations in VT_{10} which affect width, and in VT_{11} which affect amplitude. Stabilization of the supply voltage diminishes these below a significant level.

(c) Temperature dependent variations of leakage current in VT_{10} and VT_{11} which are eliminated by using silicon transistors.

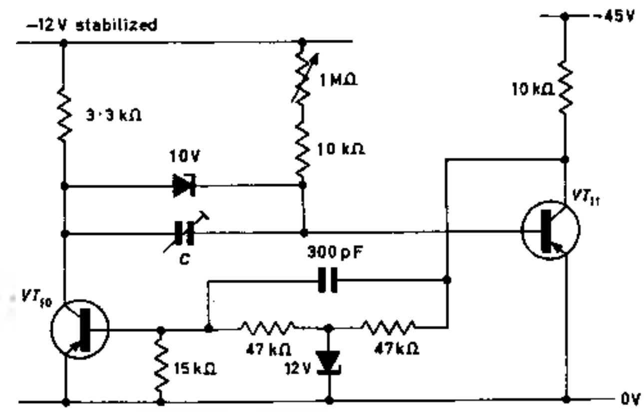


Fig. 9. An alternative pulse generator which permits a wide range of continuous width control

A pulse width of 1sec is provided using a 1·59μF capacitor.

(d) Base-emitter voltage variation in VT_{11} which depends upon temperature and supply voltage. The existence of a finite base-emitter voltage gives a deviation of the timing period from that given by the ideal timing equation where the period $T = RC\ln 2$. This variation of pulse width with change in V_{be} therefore becomes more significant at lower supply voltages. Not only must V_{be} remain constant but it should also be a constant fraction of the supply voltage.

The circuit of Fig. 7(b) isolates the timing capacitor from the collector of VT_{10} by a diode. It can be shown that when the forward voltage across the diode equals V_{be} of VT_{11} the timing period is independent of supply voltage. Matching of the temperature coefficients of the two junctions also gives temperature stability. Since the forward voltage temperature coefficients of semiconductor junctions is close to $-2\text{mV}/°\text{C}$ even random selection of the diode gives considerable temperature and voltage independence of pulse width.

Pulse amplitude is from V_{sat} of VT_{11} (about 0·1V) to a value nearly equal to

$$\left[\frac{R_3}{R_2 + R_3}\right] 45\text{V}$$

Even if the supply to VT_{11} is not stabilized the variation of V_{sat} is of little consequence in relation to a pulse of about 40V amplitude. Thus the only variable of consequence is that of pulse amplitude with supply voltage. In the final design the output of the pulse generator is 93 per cent of supply voltage. A Zener diode of about 35V across VT_{11} will stabilize pulse amplitude against supply voltage variation down to 38V.

The reverse base-emitter potential on VT_{11} is approximately equal to the stabilized line voltage, i.e. 12V, and the maximum collector-base voltage is approximately the sum of 12V and 45V. The choice of devices in small cans with these breakdown characteristics is surprisingly small and proved to be a significant design limitation.

Spacer

This is a further delay circuit designed to separate the two coupled pulses. As a wide range of continuous control is desirable the circuit used is one described by Bannister[3] and shown in Fig. 8.

When VT_7 is saturated it derives its base current mainly from the pathway $R_1 R_2 MRZ$ and the value of RV_3 is therefore not critical. When a positive step voltage is applied to VT_6 collector the regenerative switch across the timing capacitor C takes the base of VT_7 to a potential of approximately $+12$V. The capacitor C then rapidly discharges through $R_1 R_2 MRZ$ to a potential of $+V_z$ when the Zener cuts off and discharge continues through RV_3 with a time-constant of RV_3C. If V_z is selected to be 20 per cent below the supply voltage V then design conditions are readily met.

The value of R_2 is obtained from the formula

$$V/R_4 = h_{FE}\,\frac{V - V_z}{R_1 + R_2}$$

where R_1 and R_4 are normally equal. This ensures that VT_7 saturates. The variable RV_3 should be in series with a fixed resistor to limit the base current to a safe value and to permit switching to occur on the trigger pulse available. Values less than $3R_4$ should be avoided. The timing capacitor switch should have an off position which applies a bias to the base of VT_7 to ensure that it is continually saturated. An 82kΩ resistor from the -12V line does this.

The spacer output from the collector of VT_7 is again

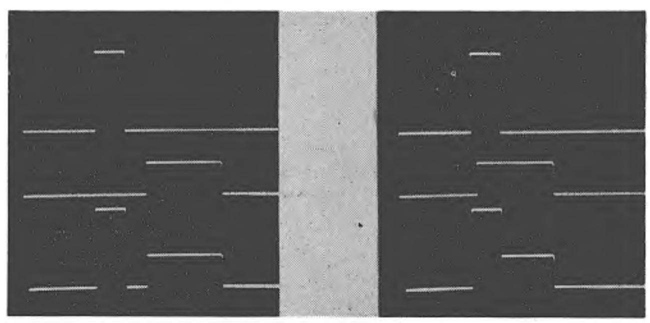


Fig. 10(a) (left). The output of the first pulse generator (uppermost trace) and the output of the second pulse generator (middle trace) are combined in the output stage (lowermost trace)

Fig. 10(b) (right). The two pulses now overlap but are combined in the output stage without summation of amplitude

a negative going rectilinear pulse, after differentiation the trailing edge drives the second pulse generator.

Alternative Pulse Generator

When pulse durations greater than 20msec are required the alternative pulse generator in Fig. 9 may be substituted. This is similar to the spacer unit and provides pulse durations up to 1sec without using electrolytic capacitors. In addition this circuit permits a large range of continuous and linear width control, the maximum/

minimum width ratio being 100 in this circuit.

Although this pulse generator does not possess the considerable temperature stability of the other version the variation in pulse width is still less than 1 per cent for a 25°C change.

This circuit tolerates supply voltage variation within the limits of 48 and 35V with a pulse width change of less than 0·5 per cent.

The output pulse has an amplitude equal to $[V_z + (47/57)(V - Z_t)]$ or approximately 0·82V + 2, i.e., 39V for a 45V supply.

Power Supply

The advantages of dry battery operation have already been discussed. The disadvantages for this design are:

(a) Pulse amplitude is a function of supply voltage.

(b) Battery internal resistance is high and must be decoupled.

(c) The necessity to check and replace batteries.

(d) The expense of battery replacement.

The stabilized −12V line is designed to function on a supply voltage down to 35V and consumes from 13mA at 45V to 9mA at 35V. The current consumption from the unstabilized −45V line is 9mA in the quiescent state rising to 55mA during an output pulse of 40V. With a 100Ω load across the output this figure is increased by 400mA.

Thus the overall current consumption depends significantly upon the mark-to-space ratio of output and to a lesser extent upon load resistance. It is difficult to foresee a use for the stimulator which will give rise to an average current drain of more than 25mA. The life of the B.104 battery to a 35V end-point is then at least 60 hours.

Battery aging can give rise to an output pulse amplitude drift of 0·1V/h, a figure of very little significance in dealing with biological material. Battery voltage may be monitored by a 50V meter on the control panel.

The decoupling of battery internal resistance is discussed in relation to output impedance below.

The component cost of providing a mains stabilized supply to each stimulator (say £7) represents the expense of supplying replacement batteries for average use over seven years in the author's department.

Output Stage

The output from each pulse generator is passed through a cascade of emitter-followers to lower the impedance and provide means of attenuation. It is convenient to make the first attenuator a range selector with values of 40, 20, 10, 5, 2 and 1V. The second attenuator may be stepped or continuous according to preference. There is some alinearity of the lower ranges due to the additive effect of the base-emitter voltages of the emitter-followers. The most common fault in commercially available stimulators is the siting of the pulse attenuator. When this is placed in the final stage the output impedance rises as the pulse is attenuated. With low resistance loads this produces a most alinear control over output voltage and this necessitates constant monitoring of the stimulus on an oscilloscope.

The two pulses are combined through emitter-followers so that there is no interaction of the attenuator controls and so that temporal overlapping of the pulses does not produce summation of amplitude (Fig. 10).

Output Impedance

The fourth emitter-follower VT_{20} is a power type with a peak current rating I_{cm} of 10A and an average rating $I_{c(av)}$ (20msec) of 6A.

As external loads can be as low as 40Ω the stimulating current may be 1A. This alone does not require a power transistor but this current would cause considerable junction temperature rise in a low power device. The value of the power device lies in protection against accidental short-circuit of the output terminals.

The output impedance of the stimulator is limited to the impedance of the power supply which for the B.104 battery is about 50Ω. A 2 000μF decoupling capacitor gives an effective output impedance of less than 5Ω with a decoupling time-constant of 200msec for a 100Ω load. This is ten times the longest pulse of 20msec and gives negligible pulse droop.

The limitations of the decoupling capacitor are again related to the prevention of damage if the output terminals are short-circuited. The peak short-circuit current is 8A (40V across 5Ω) and as the time-constant is now 10msec the average over 20msec is about 4A. These are both within the limitations of the OC36 but the decoupling capacitor must not be increased above 2 000μF unless the output impedance is raised.

The leakage current of the OC36 is typically 100μA and with reverse bias may be appreciably less. This is shared between the internal emitter resistor of the OC36 and the external load. As the biological tissue is commonly surrounded by low resistance electrolytic solution which represents an appreciable shunt, the effect on the tissue is of little consequence. However polarization of the electrodes can be troublesome. Increased junction temperature due to short-circuit of the output stage can lead to a much higher leakage current and overall the low leakage characteristic of a silicon junction is desirable.

Because of the non-availability of suitable power pnp silicon transistors this is achieved by placing a high current silicon diode in series with the output. The larger V_{be} characteristic of this junction makes the alinearity of the attenuator controls more noticeable in the 1 and 2V ranges although tempering of the attenuator resistor chain compensates to a large extent.

Output Capacitance to Ground

Output capacitance to earth depends partly upon the casing of the stimulator. A metal case is really necessary to prevent interference with high gain amplifiers etc. This case should be earthed and this will increase the capacitative link between the output and earth. If the adjacent amplifiers are adequately screened, a non-metallic panel and case considerably lower capacitance. Values of 30pF are obtainable in this way but care is necessary in coupling the output of the oscilloscope synchronizing unit if this value is to be maintained.

If it is necessary to balance the stimulus output about earth this may be done using the potentiometer across the output shown in Fig. 5. The control panel should provide a means of disconnecting the potentiometer slider from earth for normal use with the stimulus floating.

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SHORT NEWS ITEMS

The third part of BS 3939: *Graphical symbols for electrical power telecommunications and electronics diagrams* has now been published. This part contains Sections 7 and 8 containing symbols for rotating machines and transducers respectively, together with a revision of the contents list of the general introduction.

Further parts will be published extending the scope of the standard.

Copies of the above standard BS 3939: Sections 7 and 8 may be obtained from the BSI Sales Branch, Newton House, 101-113 Pentonville Road, London, N.1. Price 8/- each (postage 9d extra to non-subscribers).

An extensive programme of development in computer work, centred on a newly installed ICT 1905 machine with a 32k word core store and magnetic tapes, has been launched by the University of Strathclyde, Glasgow.

A Department of Computer Science has been created, and the University has advertised recently for a Professor to take charge of its development. Other academic staff are being recruited, and the technical team of 12 who will operate the 1905 has already been formed.

As part of the present expansion it is hoped to introduce soon an undergraduate degree course in computer science.

The Department of Computer Science will be responsible for providing a centralized computer service for Strathclyde. It will take over and develop existing teaching and research facilities in the computer field from the Department of Mathematics, although numerical analysis will remain the province of the latter under Professor David Butler. At the same time the Department will provide a pool of computer 'know-how', and its members will be available to discuss new projects with University staff and others. In this way the exchange of ideas developed by computer users working in different fields will be promoted.

At Strathclyde there are already some post-graduate courses in computer science, and a basic programming course is an integral part of the studies of all undergraduate students in engineering and mathematics. There is also a major option in Numerical Methods and Computing in the final year of the mathematics Honours degree course.

Elliott-Automation Ltd has been awarded a contract, valued at nearly £1M for automated test equipment in support of the RAF HS.801 maritime aircraft programme. Comparable equipment in the USA costs nearly three times as much.

This new range of C700 automatic test equipment, the first outside America to incorporate a digital computer, is many times faster and more accurate than anything marketed so far. Based on the Elliott 900 series of small, low-cost computers, it will check-out a wider range of equipment, including radio, radar and inertial navigational systems. Computer-controlled automatic test equipment of this type will be essential to test the equipment fitted in supersonic transports and the next generation of civil and military aircraft.

In the HS.801 programme the C700 will test automatically more than 50 separate 'black boxes'. Simply by the addition of further programmes, the equipment will be able to provide a system of automatic testing for the thousands of electronic and instrumentation components in a fleet of civil or military aircraft.

The C700 also has another interesting feature, an electronic information display in which the computer generates written messages on a cathode-ray tube to give the operator explicit test results and to convey detailed instructions in an immediately comprehensible form.

A new British Standard recently published as BS 4119 *Method of measurement of current noise generated in fixed resistors* describes a method of measurement and associated test conditions which apply to all classes of fixed resistors used in circuits where current noise is critical. The method is not intended as a general specification requirement and limiting values of the magnitude of current noise for acceptable performance are not specified.

Copies of the above standard may be obtained from the Sales Department, Newton House, 101-113 Pentonville Road, London, N.1. Price 12/- each (postage 1/- extra to non-subscribers).

The Canadian Department of Defence Production, Ottawa, has announced the award of a production contract worth approximately £600 000 to Ferranti-Packard Electric Ltd of Toronto, for the supply of ISIS 'N' (Integrated Strike and Interception System) gunsights for the Royal Canadian Air Force CF-5 fighter. The production of the equipment will be shared between Ferranti-Packard Ltd and Ferranti Ltd, Edinburgh, Scotland. Deliveries of the first production units are scheduled for late 1967.

ISIS 'N' provides accurate aiming of guns, bombs and rockets for ground attack and air-to-air combat, and is one of the ISIS family of weapon aiming systems developed by the Ferranti organization in Edinburgh.

The National Cash Register Company (Manufacturing) Ltd, Dundee is to create three research fellowships in the Department of Mathematics of Queen's College, Dundee, when the College becomes the University of Dundee on 1 August, 1967. Each Fellowship is to be worth £2 500 per annum and the Company has also agreed to cover further incidental expenses involved in mounting and continuing the Fellowship scheme. The subjects to be pursued by the Fellows, who will be university graduates, with industrial experience in some cases, are to be mutually agreed between NCR and the University's Professors of Mathematics.

The general area of research will be in the fields of numerical analysis and computational science which are important for the rapidly expanding computer manufacturing interests of NCR and for academic developments in the University. In particular, the Fellows will cover such subjects as operations evaluation topics in the areas of simulation and analytical techniques in business data processing, design automation topics such as the computer aided design of compilers, and classical problems in mathematics for advanced development in areas such as heat transfer, optics and fluid mechanics. This close co-operation in the selection of the subjects will help ensure that research is aimed at clear-cut targets likely to be of practical value in the development of existing products or in the creation of new products and information processing systems. NCR has also undertaken to provide any special equipment or apparatus required in these research projects and the Fellows will be free to publish the results of their work.

The electrical and allied manufacturing industry reached its target of over £400M worth of exports last year. Announcing this, the British Electrical and Allied Manufacturers' Association stated that exports last year at £405M represented an advance of 7 per cent on the 1965 figures, an achievement slightly above the average export increase for all manufacturing industries taken together.

BEAMA pointed out that great efforts would be needed to maintain this rate of growth in the face of intensive world competition, and the Association plans to help member firms in their export drive by organizing a number of overseas promotions this year. The Association is also producing an Export Directory with a Buyer's Guide in five foreign languages for distribution to 20 000 overseas buyers.

Hilger & Watts International Ltd has now completed negotiations with Stankoimport Moscow for a contract worth £240 000 for precision measuring instruments for the engineering industry. This contract is the largest single order placed with Hilger & Watts for such products. The instruments are of a type also produced in Eastern Europe.

Advance Electronics Ltd and the Data Technology Corporation of America have signed a further manufacturing and marketing agreement which enables Advance to manufacture and supply the Data Technology range of modular logic cards in the UK and EFTA.

Up to 108 of these cards can be mounted in 3½in of rack space. The cards feature replaceable integrated circuits, a 5V common power supply voltage throughout and the use of nand logic which optimizes systems design. Two ranges of these cards give a performance stability in the temperature ranges of 0 to 50°C or 0 to 70°C and all circuits have 5MHz capability.

The full range of logic cards immediately available from Advance includes varied input nand cards, power nands, flip-flops, logic function cards, general circuits, crystal and RC clocks, level converters, and various output drivers. A booklet is available which describes the many applications of these cards and gives details of dimensions and other mounting considerations. An applications information manual will also be available.

A four-nation European consortium led by Elliott-Automation is proposing to Eurocontrol, the international upper airspace traffic control organization for Western Europe, an advanced computer data handling system to pave the way for automatic air traffic control in the 1970's. If the bid is successful, the equipment is first to be tested in Eurocontrol's new experimental centre at Bretigny, near Paris, and will assist Eurocontrol in formulating their future operational air traffic control centre requirements.

The other members of the consortium are Ateliers de Constructions Electriques de Charleroi (ACEC) in Belgium, Standard Elektrik Lorenz (SEL) in Germany and Compagnie Generale de Constructions Telephoniques (CGCT) in France.

These companies are already working as a team to produce for Eurocontrol an advanced automatic secondary radar installation controlled by an Elliott 900 Series computer, capable of managing a complete air traffic sector without the usual primary surveillance radar coverage. The new data handling system now being proposed to Eurocontrol is an advanced development of this technique, making use of the established experience of all the team members in special equipment for air traffic control. It will allow Eurocontrol to investigate multiple computer installations, display techniques, control procedures and human factors of an operational system.

Elliotts are leading the proposed new development with system design, the computer complex and the programmes. ACEC contribute display controls, including keyboards, rolling ball, light pen, drive logic and power supplies. SEL contribute the advanced radar extraction and display system, data links and interfaces with existing equipment. CGCT provide consoles, telecommunications installation including a computer controlled network; and they would be responsible for installation at Bretigny.

A conference on frequency generation and control for radio systems will be held at the Institution of Electrical Engineers, Savoy Place, from 22 to 24 May 1967.

The aim of the conference is to review the developments in techniques of frequency generation, synthesis, control and distribution as applied to radio communications, navigation and radar systems.

The programme of the conference will be:

(1) Oscillators
- (a) atomic and transferred electron sources
- (b) crystal controlled oscillators
- (c) measurement and correction

(2) Control and derivation of frequency
- (a) synthesizers
- (b) phase and frequency locking and tracking systems

(3) Applications.

The Conference is being sponsored by the Electronics Division of the IEE in collaboration with the Institution of Electronic and Radio Engineers and the United Kingdom and Republic of Ireland Section of the Institute of Electrical and Electronic Engineers.

Registration forms and further details are available from the Conference Section, IEE, Savoy Place, London, W.C.2.

The first International Broadcasting Convention is to be held from 20 September to 22 September 1967 inclusive at the Royal Lancaster Hotel, Lancaster Gate, London, W2.

The Convention, organized jointly by the Electronic Engineering Association and the Royal Television Society, will comprise an exhibition of broadcasting equipment by leading manufacturers and a conference at which delegates can attend lectures by leading experts in all aspects of broadcasting.

Cable and Wireless Ltd and the Canadian Overseas Telecommunication Corporation have announced their intention to provide a jointly-owned coaxial telephone cable between Bermuda and Canada.

The new cable—approximately 800 miles long—will have 480 circuits and a maximum capability of 640 telephone circuits. Repeaters, used throughout the cable's length to maintain signal strength, will be fully transistorized. This will be the largest capacity and first transistorized telephone cable to be laid and owned by Cable and Wireless Ltd and C.O.T.C.

The 'CANBER' cable will be the third telephone cable radiating from Bermuda. It will provide a second route to the North American continent and will also interconnect with the Bermuda/Tortola cable southwards. The latter connects with Florida by Telephone cable and also with the recently completed Eastern Caribbean Tropospheric Scatter Radio system, which is to be extended to Guyana.

The 'CANBER' cable will land in Canada in the vicinity of Mill Village, Nova Scotia, and will permit interconnection with the nearby Canadian earth station.

Redifon Ltd, a member of the Rediffusion Group, engaged in the manufacture of flight simulators and telecommunications equipment, announces that it has completed negotiations for the acquisition of all the shares of General Precision Systems Ltd of Aylesbury, a company similarly engaged in flight simulator manufacture. General Precision Systems Ltd is a subsidiary of General Precision Equipment Corporation, U.S.A. The Air Traffic Management Division of G.P.S. is excluded from this transaction.

A new design in party line telephone systems—the Identifone—has been developed by the Plessey Automation Group's Data Handling Division.

The Identifone system is suitable for numerous applications including signal post telephones for railways, motorways, and emergency call telephones where identification of location is necessary, such as police and fire stations, mines and pipeline checkpoints. The system can also be integrated into automatic telephone networks where calling-line identification is required.

The Identifone offers complete security since the master station is called by a unique signal, and once contact is established the conversation becomes secret. Power supply for signalling and speech is fed from the master station along the line.

Optional extras of the Plessey Identifone system include automatic routing and selective call back. Electronic circuits permit the operation of special party line telephones over cables of smaller gauge than possible with older systems. For example, on railway signal post telephone systems 20lb/mile cable may be used where 40lb/mile cable has been necessary before. The financial saving in cable may often offset the whole cost of replacing existing systems by Identifone.

The C.E.I. examination will form the subject of a one-day conference to be held on Tuesday, 11 July (10 a.m. to 4 p.m.), at the Institution of Civil Engineers

This new examination of the Council of Engineering Institutions is being introduced in October 1967 and will shortly replace the current separate examinations of Institutions as the academic qualifications required by chartered engineers. The purpose of the conference is to provide a forum for discussion on the new examination and it is hoped that this will be particularly useful to senior staff of colleges who will be considering what adjustments may be needed to their existing range of courses, and to senior managers and training officers in industry who will be concerned with the supply of professional engineers and the training of junior members of staff.

Speakers will include eminent engineers and academics who have been closely concerned with the formulation of the new examination and there will be the opportunity to obtain amplifying information from them. Particulars of Part 1 and Part 2 of the examination have already been published in C.E.I. Statements Nos. 1-4, and the policy for exemption from the examination is now being published in Statement No. 5. The subject of 'The Engineer in Society', which is a compulsory subject of Part 2, has created especial interest and will receive particular attention at the conference. To stimulate discussion there will be contributions by a principal of a college of technology and a prominent industrialist. The presidents of some of the constituent institutes will be participating.

Registration forms giving details of the conference will be available after 1 May from the Secretary of the Institution of Civil Engineers, Great George Street, London, S.W.1.

Explosive forming, friction welding, laser holography and computer aids to engineering design are among many advanced physical techniques being studied in Government research laboratories. To stimulate industry into making use of this work the Institute of Physics and the Physical Society in collaboration with the Ministry of Technology have organized a conference, 'Industrial Physics—the Contribution of Government Laboratories'. It is to be held in Harrogate, Yorkshire, on 7 to 9 June, 1967.

Representatives from industry who attend the conference will be told of fifteen techniques, including the four above, which could profitably be used in industry to improve production methods and productive efficiency. These are based on work being carried out in laboratories of the United Kingdom Atomic Energy Authority, the Science Research Council, Ministry of Defence and Ministry of Technology.

Three papers at the end of the conference by members of the Ministry of Technology, Central Electricity Generating Board and the National Research Development Corporation will cover the exploitation of these advanced techniques by industry. A large part of the conference will be given over to discussion of the ideas presented in these papers.

Application forms for advance registration for attendance at the conference can be obtained from: The Meetings Officer, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

Qatar, one of the wealthy oil states in the Arabian Gulf, has started to establish its first broadcasting station. The project was put to internation tender, and the contract, worth approximately £200 000 has been won by the Marconi Co. Ltd; they have already started to supply the equipment, and it is anticipated that transmissions will begin early next year.

A 100kW short wave transmitter will be used to provide a coverage over a 1 200 miles radius, and for local broadcasting this will be supplemented by a 10kW medium wave transmitter. Both these transmitters, broadcasting identical programmes, will be installed in a broadcasting station a few miles outside Doha, the capital of Qatar. The station will be connected by links to a Marconi equipped Studio Centre, situated in Doha itself.

The installation is one of a series of comprehensive modernisation projects which were initiated by the Government of Qatar a number of years ago. The object is to promote the living standards and prosperity of the Qatar people and to this end, economic, social, cultural and administrative improvements are being made.

Qatar will be the tenth country in the Middle East using Marconi broadcasting equipment.

The associations of active electronic component manufacturers in Belgium, France, Germany, Italy, Netherlands and the United Kingdom confirmed their preliminary constitution at a meeting in Munich in January 1967, and formed a European Committee under the name of C.E.M.A.C.—Committee of European Associations of Manufacturers of Active Electronic Components.

This new link-up of the European associations is the result of a series of liaison meetings which have taken place over the past year, following upon informal discussions firstly between VASCA (Electronic Valve and Semiconductor Manufacturers' Association) and SITELESC (Syndicat des Industries de Tubes Electroniques et Semiconducteurs), beginning in May 1965.

The work is directed towards the common interests of European countries, especially the promotion of trade by, for example, the elimination of technical barriers.

The first objective is to promote the compatibility of specifications, organizations and procedures for active electronic components throughout Europe, and arrive at compatible national specifications, which will be mutually recognized in all European countries and, through I.E.C. (International Electrotechnical Commission) developed into world standards.

The procedures and specifications being determined are directed towards all users of active electronic components and are intended to reduce the varieties of specifications which exist at present, improve the quality assurance and aid the work of I.E.C. national committees.

C.E.M.A.C. is very similar to C.E.P.E.C. (Committee of European Associations of Manufacturers of Passive Electronic Components) and liaison has been organized to ensure close cooperation on matters of common interest.

The associations concerned are FABRIMETAL (Belgium), SITELESC (France), ZVEI (Germany), ANIE (Italy), FOEGIN (Netherlands), and VASCA (United Kingdom).

The president of C.E.M.A.C. is Mr. P. H. Spagnoletti of VASCA and the vice-president, Mr. F. Dumat of SITELESC. The secretary of C.E.M.A.C. is Mr. P. A. Fleming of VASCA, Mappin House, 156/162 Oxford Street, London, W.1.

LETTERS TO THE EDITOR

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

A Modified Cascode Circuit Using Complementary Transistors

DEAR SIR,—Although planar manufacturing techniques have facilitated the construction of transistors in which the collector-base feedback capacitance is reduced to a low value, and this can be as low as 0·2pF in the most carefully designed units, nevertheless in the majority of small-signal r.f. amplifier transistors this capacitance will be in the range 0·6 to 2pF. This order of feedback capacitance will lead to instability in high gain tuned r.f. amplifier stages unless some form of feedback neutralization is employed. In the case of a conventional i.f. amplifier, this will take the form shown in Fig. 1. This arrangement imposes some limitations

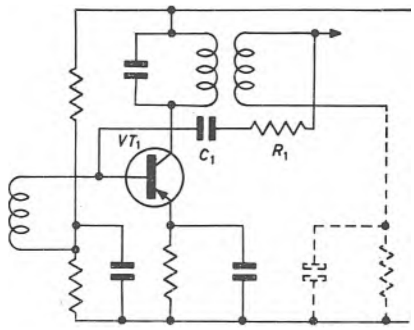


Fig. 1. Neutralization with conventional amplifier

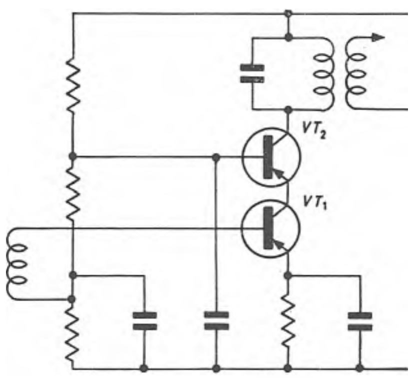


Fig. 2. Cascode amplifier

on the type of interstage coupling which may be employed and requires careful choice of the value of the unilateralization components C_1 and R_1.

However, with the growing availability of low cost silicon planar transistors, the use of a cascode amplifier of the form shown in Fig. 2 no longer imposes too heavy a cost penalty. Because of the effective internal screening of transistor VT_2, the feedback capacitance in this type of circuit is much reduced; overall reverse transfer admittances (Y_{12}) less than 0·001mΩ^{-1} have been quoted for this configuration; and tuned r.f. amplifiers of this type can operate stably without unilateralization.

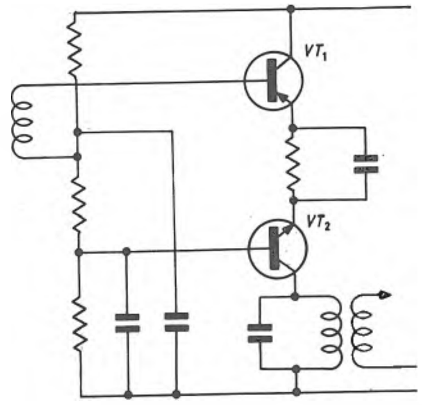


Fig. 3. Rearrangement using complementary transistors

This circuit can rearranged in the form shown in Fig. 3, using complementary transistors, so that transistor VT_1 is employed as an emitter-follower. This offers the advantage of a higher input impedance, as well as improved h.f. performance from the input transistor. Because the collector and emitter currents of VT_1 are substantially identical the gain of this arrangement is not significantly different in other respects from that of the arrangement shown in Fig. 2.

Yours faithfully,

J. L. LINSLEY HOOD,
Taunton, Somerset.

Augmented D.C. Follower

DEAR SIR,—A unity gain d.c.-coupled amplifier, having a high input impedance (d.c. and a.c.) and low offset and drift, finds application in the fields of measurement and instrumentation. Fig. 1 shows the arrangement of one approach to the design of such an amplifier developed in this laboratory. F is an impedance-transforming, level-shifting, input stage of gain G_1 the purpose of which is to supply differential drive to the inputs of the high gain integrated amplifier A of gain G_2. Obviously, from feedback considerations,

$\Delta V_o = \Delta V_I G_1 G_2/(1 + G_1 G_2) \simeq \Delta V_I$ if $G_1 G_2 \gg 1$.

The way this arrangement is practically implemented is shown in Fig. 2. VT_{1a}, VT_{1b} (Amelco 2N3934) constitute an integrated pair N-channel depletion-mode junction f.e.t.; a matched m.o.s. device could equally well be used. Matched bipolar pair VT_{2a}, VT_{2b} (SL301-A) and their associated components R_{2a}, R_{2b}, MRZ_1, R_3, give constant d.c. source current (0·2mA) bias for VT_{1a}, VT_{1b}. R_{1a}, R_{1b}, a few kΩ in magnitude, and calculated from the measured mutual characteristics, are necessary in order that the differential inputs to A may be approximately zero volts for the condition $V_o = V_I = 0$V.

R_{1b} is made variable to permit fine adjustment of this condition. Because of the circuit technique and f.e.t. used $(d/dT)(V_{GS1} - V_{GS2}) \simeq$ few $\mu V/{}^\circ$C. R_{2b} may be made variable to permit unequal currents in VT_{1a}, VT_{1b}, and thus allow the overall drift of the complete unit to be set to a minimum. However, two points should be borne in mind: (1) this does mean the complication of two inter-dependent adjustments; (2) because of gate leakage currents the temperature

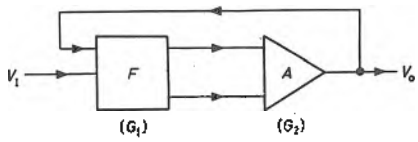
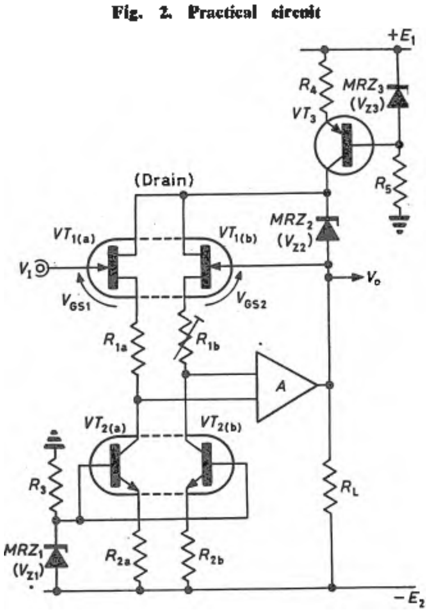


Fig. 1. Arrangement of amplifier

Fig. 2. Practical circuit



dependence of the unit depends also on the output resistance (perhaps unknown) of the circuit being investigated and cancellation of effects due to this requires a resistance of equal magnitude connected between the amplifier output and the gate of VT_{1b}.

Thus VT_{1a}, VT_{1b} should have low gate leakage currents. VT_{1a}, VT_{1b}, are each operated at constant V_{DS}, as well as I_{DS}, irrespective of input signal swing (providing no over-loading occurs). This arises because of the presence of Zener diode MRZ_2 driven by a current source formed by VT_3, MRZ_3, R_4, R_5.

The l.f. gain from the input of VT_{1a} to the input of A is

$\simeq \mu R_{in}/\{r_{DS}+(\mu+1)(R_{1a}+R_{in})\} \simeq 1$

where R_{in} = input impedance of each side of A($>$ 100kΩ) and r_{DS}, μ, have their usual significance for f.e.t.'s. A similar relationship holds for VT_{1b}.

Hence the loop gain behaviour is

determined by the characteristics of A, the performance of which is well determined and reproducible. The bootstrapping of the drains reduces the capacitive loading of the unit and ensures the maximum bandwidth for a given source resistance. MRZ_1, MRZ_2 must be chosen so that VT_{1a}, VT_{1b}, VT_2 do not saturate for extremes of input signal variation.

If the input connexions to A are reversed, the feedback disconnected, the gate of VT_{1b} earthed and the drains of VT_{1a}, VT_{1b} are taken to $+E_1$, the unit may be used in operational amplifier applications.

B. L. Hart,
West Ham College of Technology.

Phase Inversion in an Overloaded Amplifier

Dear Sir,—In connexion with Phase Inversion in an Overloaded Amplifier[1] I should, perhaps, have extended my previous letter[2] to comment on the application of the circuit as a frequency multiplier for triangular waves.

It is evident that if a triangular wave, of amplitude V, is applied to the input then a similar waveform of half the amplitude but twice the frequency is generated at the output. Such circuits may, with suitable biasing, be fed in series to multiply by a power of 2 as shown in Fig. 1, where $R_2 \gg R_1 \gg R_{source}$, and junction voltage drops are neglected.

The voltage transfer characteristic for a single stage is shown by the solid line in Fig. 2. The addition of a resistor between base and collector (as shown in Fig. 3) yields a transfer characteristic of the form shown by the broken line which it approaches for large γ, thus permitting frequency multiplication by 3 as shown in Fig. 3. The output peaks are of amplitude $(V/2).(\gamma/2(1+2\gamma))$ and $(V/2).(\gamma/2(1+\gamma))$, they occur when the input amplitude is $(V/2).(\gamma/1+2\gamma)$ and V.

A sinusoidal waveform when applied to one stage of Fig. 1 yields an inverted sine squared waveform and when applied to Fig. 3 a quasi-triangular waveform is produced consisting of sinusoidal arcs.

Yours faithfully,

D. J. Grover,
Marconi Instruments Ltd.

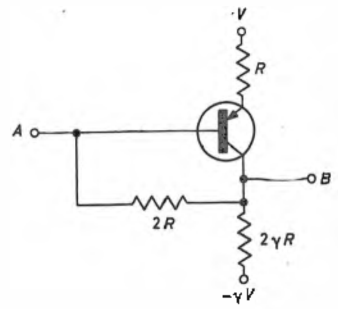
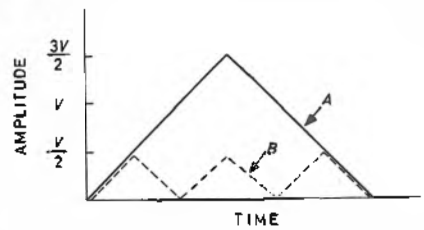


Fig. 3. Modification for multiplication by three



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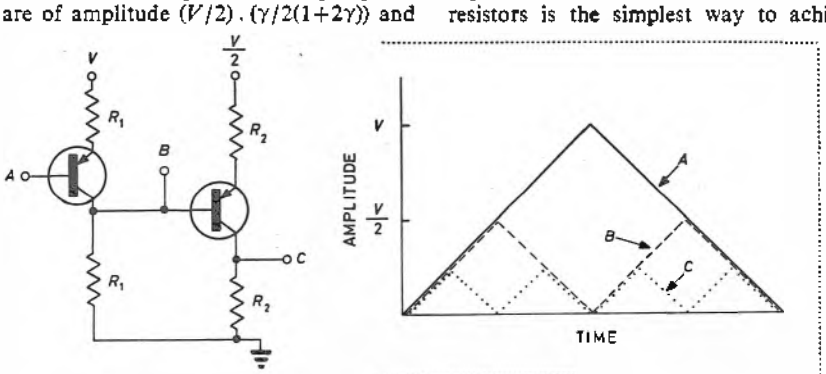
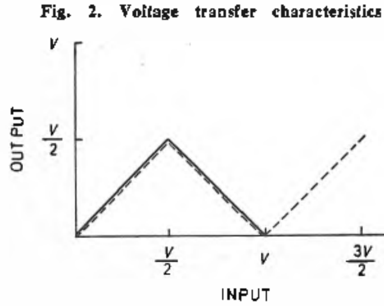


Fig. 1. Multiplier circuit and waveform

Fig. 2. Voltage transfer characteristics



Economy and Reliability in Nucleonic Instrument Design

Dear Sir,—I would like to support the recent comments made by Mr. B. Shepherd (February issue) on how to achieve reliability and economy in instrument design. We here are faced frequently with the development of industrial electronic equipment and we have found that the use of epoxy encapsulated silicon planar transistors together with metal film or metal oxide resistors is the simplest way to achieve reliability and economy in our environmental conditions. We estimate that the extra component cost is saved many times over by the reduced development time that is necessary, and the reduction in commissioning (debugging). For some time we have regarded equipment using germanium transistors as second rate for our purposes because of the difficulty of ensuring adequate safety margins against temperature effects.

We appreciate fully the equipment manufacturer's difficulties at the moment when the British semiconductor manufacturers do not have an adequate range, and have to admit that a large proportion of our semiconductors are obtained from abroad. However, I would suggest that the improved performance of planar devices will allow the production of a restricted range of transistors for industrial use having limited interchangeability between sources along the lines that evolved 20 years ago with the valve types used in domestic radio sets.

Yours faithfully,

D. Shaw,
The United Steel Companies Ltd.

The author replies:

Dear Sir,—Mr. Shaw's comments are very welcome. British semiconductor manufacturers, in general, tend to lag some two years behind their American counterparts. Although we have the ability to produce these components we still lack the low cost, reliable, commercial ranges.

As an example I would like to cite an instrument which is at present being designed and will be used by medical and research personnel. It contains approximately 150 integrated circuits and a further 150 transistors. The I.C.'s chosen were a DTL range originally specified as Fairchild. Two manufacturers in this country now manufacture and market this range under licence. In October last year Motorola introduced the same range, with the same pin configuration, in a plastic package, at a price less than half that of its nearest British competitor. Other U.S. stockists offer different ranges at similar prices.

The excuses that they have been subsidized by the U.S. military and space programmes or that our quantity requirements are smaller, are of little use to the instrument designer, whose instruments must, nearly always, compete with imported instruments. This situation is aggravated in our case as 40 per cent of our instruments are exported.

I have a natural reluctance to obtain components from U.S. stockists, but I have a duty as a designer to my employers. Having sent letters to seven suppliers for an I.C. gated sense amplifier, of which I require 20 per instrument, I was quoted prices varying from £9 to £41 each. The imported device I am using now costs £4.

Yours faithfully,

B. Shepherd,
Isotope Developments Ltd.

NEW BOOKS

Basic Automatic Control Theory

By G. J. Murphy. 832 pp. Med. 8vo. Van Nostrand Ltd. 1966. Price £5 18s.

THIS introductory book is a revised edition of the original text published in 1957. The new material includes a more detailed discussion of stability (including Liapunov Methods) and new chapters on state representation, carrier control systems, sampled data systems, probability and random processes and optimum synthesis.

Some of the material in the first edition, such as the chapter on analogue computation has been discarded in favour of the additional topics and this change has resulted in a more up-to-date text. As the author admits himself, perhaps a more severe pruning could have been carried out since the book is now rather bulky (some 832 pages) as an introductory text. Due to the size and because the author wished to keep the text at an introductory level, certain aspects of modern control theory are omitted such as Pontryagin's Maximum Principle, dynamic programming, calculus of variations and iterative methods for boundary value problems.

The new edition follows the original text but with some revision up to chapter 5 with chapters on physical analogues, components, Laplace transforms and transfer functions and signal flow diagrams. The original chapter 6 on types of control systems is now replaced by one on the concept of state. This is followed by chapters on stability, frequency response, design and then the new chapters already mentioned. Despite the author's plea that a revision was not justified, some of the material concerned with electronic circuits in chapter 3 may appear, to the young reader, to be rather dated since the circuits described are based on thermionic valves.

In the chapter on design, although sensitivity considerations claim four pages of discussion and again a brief mention in the chapter on sampled data systems, no real use is made of these fundamental concepts in the design examples. Since these concepts describe the fundamental reasons for using feedback the reviewer feels that this aspect is the most glaring failure of this new edition.

A welcome feature is the list of references at the end of each chapter together with a varied set of examples. The latter, however, would be of more interest to the individual student reader if solutions had also been given.

In conclusion, this new edition is a significant improvement on the original text and readers who liked the latter will probably find this book very much to their liking. Despite the shortcomings mentioned it is a book worthy of a place in the control engineers' library.

R. J. A. PAUL

Transistor Physics

By K. G. Nichols and E. V. Vernon. 332 pp. Crown 8vo. Chapman & Hall Ltd. 1966. Price 40s.

THIS is a text containing a good deal of useful information on semiconductor physics and, to a lesser extent on transistor theory. The price is a very reasonable one for a book of this size and will make it attractive to students.

The total of 320 pages of text is divided as follows: Classical atomic theory, Quantum mechanics and band theory, Statistical mechanics—a total of 84 pages. Semiconductors in thermodynamic equilibrium, Current flow, Non-equilibrium conditions, Current flow in semiconductor junctions—a total of 151 pages. Minority carrier transistor—46 pages. Space-charge—limited currents in semiconductors and insulators, Majority carrier transistor—27 pages.

The reviewer's criticism is that the book's general conception falls between two stools. For a textbook on transistor physics, which it purports to be, it is clear that 46 pages devoted to the principal topic is a sadly inadequate volume. The casualties are, to name but a few notable omissions, high-injection-level theory, large signal transients, derived structures. It would have been much better to deal with these and other transistor topics at the expense of some of the details of fundamental principles.

Considered as a textbook on semiconductor physics, on the other hand, the book lacks depth and rigour, of which the most notable example is the exclusion of Boltzmann's transport equation. This forces the authors into ingenious mental acrobatics in the derivation of mobility—pp. 110-113 which at times give simply incorrect results. Yet Boltzmann's equation is certainly not more complicated or difficult than some of the other topics the authors develop in their text.

The conclusion from these criticisms is that it is extremely difficult to provide in the form of one reasonably priced book an adequate coverage of the entire theoretical solid state background and a reasonable account of transistor principles. The present volume goes part of the way to this ideal.

A. K. JONSCHER

Laser Receivers

By Monte Ross. 405 pp. Med. 8vo. John Wiley & Sons. 1966. Price 113s.

IN the book 'Laser Receivers' Monte Ross has brought together a number of topics which concern optical receivers from the point of view of the optical electronic engineer. An important part of the book (chapters 2 and 3) deals with noise and information carrying capacity of an optical system. The use of Bose-Einstein statistics for describing photon fluctuations in radiation is introduced and recent work on measurements of optical noise is briefly mentioned. The additional noise produced by photo-detectors and thermal background is dealt with in detail. One rather distracting feature which makes itself felt at this point is the use of abbreviations like NEP, BLIP and particularly IROP.

Chapters 4 and 5 are the most important in the book, and deal with receiver techniques and photo-detectors. An analysis is given of the use of mixing techniques and the improvements in signal-to-noise ratio which can be expected. Chapter 5 deals with the various types of photo-detector and their performance and limitations in practice. Chapter 6 deals with the techniques of modulating light beams and although not directly relevant to the subject matter of the book, forms nonetheless a very useful survey of this field.

Chapter 7 deals with the background noise to be expected, particularly if the field of the laser receiver includes the moon or a planetary body. Chapter 8 describes the effect of the medium on the laser beam. Turbulence phenomena and scattering are well described. A short description is also given of optical guides. Chapter 10 describes a number of possible laser systems for optical radar and communication, together with some interesting comments on meteorological studies by laser scattering. The last chapter deals all too shortly with possible applications of laser receivers and ends with the very exciting possibility of interstellar communication with aliens by laser beams.

This book should prove to be a very useful introduction for workers in the rapidly changing field of laser technology, and like all similar books it has an abundance of references, particularly to American reports.

A. J. Fox

Zur Frage der Verhütbarkeit von Kollisionen in der Binnenschiffahrt (Aspects of Collision Prevention on Inland Waterways)

By F. Schröter. Research Reports of the Land Nordrhein-Westfalen No. 1668. 114 pp. Royal 8vo. Westdeutscher Verlag, Cologne. 1966. Paperback, Price DM.61.30

INCREASING density of shipping on inland waterways requires the development of a very simple technical aid to give increased safety in navigation under conditions of bad visibility. Sketching in what has been done in neighbouring areas, e.g. navigation on the high seas and in harbours, Prof. Schröter ascertains that the whole spectrum below 8mm down to the 'optical' wavelength has to be investigated to determine conditions for simple and economical radar and direction finding equipment with suitable range and accuracy. Results show that a compromise must be sought.

The aids are not required to avoid a collision of ships already on collision course, but rather to prevent such a situation from arising in the first place. To achieve this, five proposals are made:

(1) A secondary radar presentation transmitted over a narrow band channel to any number of shipborne receivers using television techniques,
(2) Shipborne radiation of warning segments,
(3) Guide beams employing AN techniques,
(4) Beam interruption as navigational aid, and
(5) The block system used in rail traffic control.

An extensive bibliography gives a large number of English language references.

E. R. FRIEDLAENDER

An Introduction to Dynamic Programming

By O. L. R. Jacobs. 126 pp. Demy 8vo. Chapman & Hall Ltd. 1967. Price 30s.

Intended for undergraduate and postgraduate use, this monograph serves as an introduction to the subject and presents the standard procedures of dynamic programming in a logical and unified manner.

Probalistic processes are dealt with and it is shown how dynamic programming is applicable to stochastic and adaptive processes.

BBC Handbook, 1967

264 pp. Crown 8vo. BBC Publications, P.O. Box 1 AR, London, W.1. 1967. Price 7s. 6d.

Among the more technical aspects in the 1967 edition of this year book are accounts of the introduction of colour television this year, using the PAL system on 625 lines, the installation of four 250kW short wave transmitters on the Ascension Island and the development of stereophonic broadcasting in Britain.

Introducere In Cibernetica Sistemelor Discrete (Introduction to Discrete Systems Cybernetics)

By Edm. Nicolau and Alex Popovici. 272 pp. Demy 8vo. Editura Tehnica, Bucarest, Rumania 1966. Price Lei 11.50

The first of these three books, all written in Rumanian, deals in a comprehensive manner with the practical applications of modern long-distance telephony, on a national and international basis, after introducing theory and fundamental principles.

Attention is also paid to artificial satellite communication systems.

A long list of reference is given in each chapter quoting worldwide sources.

G. Rulea in this book on v.h.f. techniques discusses the applications of radio relay systems, artificial satellites, radio navigation and radio astronomy, as well as the biological and medical applications.

Included in the book are some original papers written by the author on theoretical aspects.

A list of 645 references is given quoting worldwide as well as Rumanian sources.

The third book, as its title suggests, covers the theory of cybernetics as applied to discrete systems on a mathematical basis.

The application of the digital computer is also dealt with.

Electronics From Theory Into Practice

By J. E. Fisher and H. B. Gatland. 411 pp. Crown 8vo. Pergamon Press Ltd. 1966. Price 35s.

The aim of this book is to formalize design procedures covering a wide range of electronic circuits.

An introduction is given to the thermionic valve and the transistor showing how manufacturers' data sheets are used in design calculations.

The remaining chapters each deal with specific designs and give a brief treatment of theory limited to the extraction of necessary design relationships, followed by design examples to meet given specifications.

Ultrasonic Machining of Intractable Materials

By A. I. Markov. 350 pp. Demy 8vo. Iliffe Books Ltd. 1967. Price 63s.

This book is one of the two in the general field of ultrasonic machining written by Russian authors, the other being 'Ultrasonic Cutting' by L. D. Rozenburg. Both have now been translated into English and the present book by A. I. Markov deals with the basic principles of conventional ultrasonic machining, together with vibrator design, machine tool details, technology and applications.

Telefonie La Mare Distanta (Long Distance Telephony)

By S. Coodrea et al. 508 pp. Post 4to Price Lei 30

Semiconductor Circuits: Worked Examples

By J. R. Abrahams and G. J. Pridham. 208 pp. Crown 8vo. Pergamon Press Ltd. 1966. Price 21s.

This book is the companion to the authors' 'Semiconductor Circuits: Theory, Design and Experiment', and here again the questions have been taken from a wide range of examination papers and are given fully worked solutions for the technician and undergraduate.

A Dictionary of Electronics

By S. Handel. 396 pp. Crown 8vo. Penguin Books Ltd. 1967. Price 75s.

Some 300 new terms and abbreviations are contained in the second edition of this dictionary which was originally published in 1962, and contained about 5 000 terms.

An Anthology of Philips Research 1891-1966

By H. B. G. Casimir and Ir. S. Gradstein. 469 pp. Med. 4to. Philips Technical Library. 1966. Price 37s.

The seventy-fifth anniversary of the Philips Company, Eindhoven, reached in 1966 provides the basis of this book which deals with the history, background and research progress in the company since its inception.

Physical Fields

By J. Krasnica. 183 pp. Crown 8vo. Iliffe Books Ltd. 1967. Price 17s. 6d.

First published in Czechoslovakia in 1964 and now translated into English, this book covers the classical theory and general properties of the electromagnetic field, the corpuscular nature of electromagnetic radiation and the dualism of waves and particles etc.

It also introduces advanced concepts such as eigen values and the equations of Schrodinger, Klein-Gordon and Dirac.

Transistor Specifications and Substitution Handbook

187 pp. Demy 8vo. Techpress Inc., Brownsburg, Ind. 1966. $2.95

The specifications of over 6 000 American and other types of transistors are contained in this book, together with some 18 000 substitute types.

Tehnica Frecventelor Foarte Inalte (V.H.F. Techniques)

By G. Rulea. 439 pp. Med. 8vo. Price Lei 34

Economic Computation and Economic Cybernetics Studies and Research

157 pp. Demy 8vo. 1966. The Centre of Computation and Economic Cybernetics, Bucharest, Rumania

Eight papers written by Prof. M. Manescu, President, and other members of the Centre of Economic Computation are presented in this book.

Prof M. Manescu also provides a summary.

Strahlungskopplung von Wendelantennen (Radiation coupling of helical aerials) Impedanzuntersuchungen an dielektrischen Flachantennen (Study of the impedance of dielectric plane aerials)

By Dipl.-Phys. D. Glesner and Dipl.-Phys. G. Jötten. Research Report No. 1666 of the Land Nordrhein-Westfalen. 88 pp. Royal 8vo. Westdeutscher Verlag, Cologne. 1966. Paperback, Price DM 55.80

The authors, members of the Institute for Applied Physics and Electrical Engineering at the University of the Saar, undertook this work in collaboration with the German Society for Direction Finding and Navigation. Glesner examines the radiation coupling in a group of two helical aerials and comes to the conclusion that the directional properties may be improved by suitable measure, but that this leads simultaneously to increased side lobes. To avoid this, a group of n helical aerials has to be constructed and the radiation power has to be graduated from the centre outward in accordance with a binomical distribution.

The second paper deals with examinations of the impedance of dielectric plane aerials with special reference to the flat-top type. Jötten arrives at a favourable combination of dimensions for a cylindrical disc shape and the length of a cylindrical excitation pin. Due to the design and the use of dielectrical material, this aerial is not liable to be damaged or to deteriorate under severe environmental conditions.

Correction

In the review of 'Electrical Conduction Mechanisms in Their Insulating Films' on page 189 of the March 1967 issue the author was given as A. I. Markov; it should read D. R. Lamb.

NEW EQUIPMENT

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

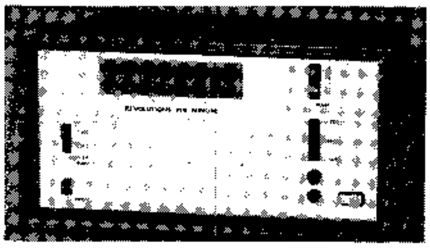
DIGITAL TACHOMETER

Smiths Industries Ltd, Industrial Instrument Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

A new four figure digital tachometer, for the accurate measurement of speed directly in revolutions per minute, has been introduced by the Industrial Instrument Division of Smiths Industries Ltd. It has been manufactured to operate from a variety of transducers including inductive and photo-electric and will accept input frequencies up to 100kHz.

Designed for the rugged requirements of industrial usage, this simple to operate digital tachometer can have a sampling time between 10msec and 10sec and extends the existing range of



tachometry already manufactured by the Division. It has a crystal oscillator reference, giving an overall accuracy of ±0·1 per cent up to 100 000 revolutions per minute. This accuracy remains unaffected by operating temperatures between 0 to 50°C.

The minimum input pulse height is 200mV peak, with a minimum width of 5msec.

EE 05 751 *for further details*

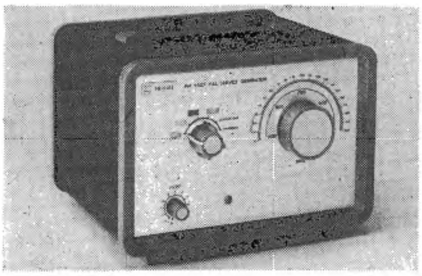
PAL PATTERN GENERATOR

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustrated above right)

A new Philips pattern generator for servicing PAL colour television receivers has been announced by the M.E.L. Equipment Co. Ltd. So far as is known it is the first which is specifically designed for the PAL system to be made available in Britain.

The generator, type PM5507, provides a PAL signal on 625 lines over a tunable range of 480 to 850MHz. The signals available are vertical lines, hori-



zontal lines, cross-hatch, dots, rainbow pattern and colour bars. The line frequency and colour sub-carrier are crystal controlled, and the picture is interlaced.

The design is completely solid state, and operation is from the mains.

EE 05 752 *for further details*

MINIATURE POWER UNIT

Sandmar Electronic Products, 68 Nina Drive, New Moston, Manchester, 10

Sandmar Electronic Products has introduced the first of its range of very low priced miniature stabilized power units for laboratory or bench use. This model has an output voltage fixed at 12V d.c. with a maximum continuous current rating of 1A. The ripple content will not exceed 1mV r.m.s. at full load. The output resistance is of the order of 20mΩ. Regulation for line and load will not exceed 0·1 per cent for changes in the mains of ±7 per cent or changes in the load from 0·1 to 1A. Current limiting short-circuit protection is incorporated, this is automatically removed when the fault is eliminated. Overall size is 3½in wide, 5in high and 7in deep.

EE 05 753 *for further details*

ELECTRONIC ENGINEERING
will occupy Stand 120 in Hall 17 at
LE SALON INTERNATIONAL DES COMPOSANTS ÉLECTRONIQUES
to be held in Paris from 5 to 10 April
and Stand 1238EF, Floor 1, at
MESUCURA
to be held in Paris from 14 to 21 April
and visitors will be welcome

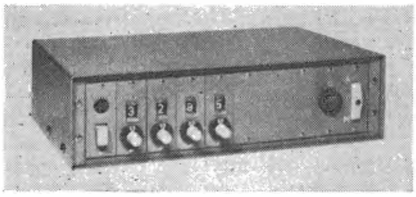
BATCH COUNTER

Radiatron, 7 Sheen Park, Richmond, Surrey

(Illustrated below)

Radiatron has developed and can now supply a fully automatic batch counter comprising four Elma single decade read-out counters and other modules. The unit can be preset to give a relay operation at any number and then either stop or automatically reset to zero and recommence counting—giving a relay operation at the end of each batch. The unit will operate direct from 220 to 240V, 50Hz.

The number to be predetermined is



set up on the rotary switches on the front plates of the counters. When the batch commences, the counters count up from zero, and on reaching the pre-set number energize relays. If the slide switch on the front is set to 'A' for Automatic, the counters will automatically reset to zero, and after a delay of approximately ½sec will start counting up again. This process will repeat itself automatically. If the slide switch is set to 'M' for Manual, the counters will stop when indicating the pre-set number. They can then be reset to zero by either pressing the push-button on the front or by switching back to 'A'. Each time the counters reach the pre-set number, a relay changes over a set of contacts for external operation (capacity 50W).

This unit is suitable for batching operations where very high speed operation is not required and expense is to be kept to the minimum. It will operate from impulse sources such as micro-switches; pressure switches; contact operations etc. as long as there is a minimum pulse length of 28msec and a minimum pulse interval of 12msec.

EE 05 754 *for further details*

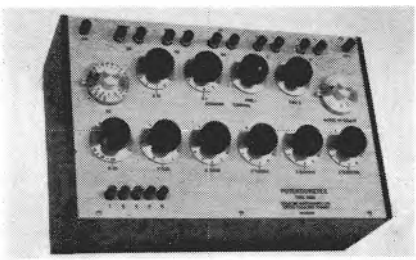
D.C. POTENTIOMETER

Distributed by: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustrated below)

Available from the Instruments Division of Claude Lyons Ltd, the Guildline type 9160G is a guarded six figure precision d.c. potentiometer with three ranges in steps of 1μV up to 2V, steps of 0·1μV up to 0·2V and steps of 0·01μV up to 0·02V. It can be used on a bench with the panel horizontal or in a rack with the panel vertical, brackets for rack-mounting being supplied. The potentiometer circuit is such that dials 1 and 2 have potential contacts only and dials 3, 4, 5 and 6 are all shunted. Contact resistance is therefore not a problem in any of the switches, and thermal e.m.f.'s are less than 0·1μV.

Accuracy specifications on the 2V range are ±(0·0005 per cent of reading + 1μV) for one year and ±(0·0015 per cent of reading + 0·1μV) for five years. On the 0·2 and 0·02V ranges it is ±(0·0015 per cent of reading + 0·1μV) for five years. The instrument incorporates complete facilities for self-checking, eliminating the need for cer-



tification by a national standards laboratory.

The standard cell dial covers the voltage range 1·017 500 to 1·019 600 in 20 steps of 100μV plus a continuously variable dial of 100μV. Standardization can be carried out on any range because the standard cell circuit is separate from the measuring circuit.

EE 05 755 *for further details*

SUB-STANDARD FREQUENCY GENERATOR

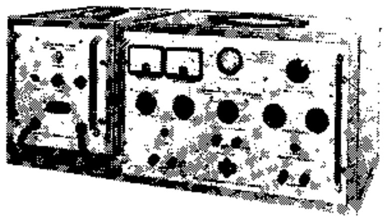
Contronics Ltd, Garth Works, Deepcut Bridge Road, Deepcut, Camberley, Surrey

(Illustrated above right)

The sub-standard frequency generator type 7C-V-30T is a development of the instrument originally designed some 15 years ago to the requirements of the Central Electricity Authority. The new equipment, employs semiconductors throughout, apart from the chronoscope tube.

The instrument gives up to 30W output continuous (60W intermittent) at seven crystal controlled fixed frequencies between 46 and 52Hz, or at variable frequencies in the range 44 to 54Hz. The variable oscillator stability being better than 0·01 per cent in 30sec.

A 'Chronoscope' (synchroscope)



cathode-ray tube is incorporated, the deflector plates of which can be switched to either a.c. from the mains, an external source or from the internal variable oscillator. The c.r.t. displays a line trace which remains stationary if the frequency of the applied a.c. is identical with the crystal generated frequency, but otherwise rotates and so enables the precise frequency difference to be measured.

The front panel is fitted with a voltmeter to monitor the output voltage and an ammeter to give a nominal indication of the power being taken by the load.

To facilitate portability, the equipment is constructed in two separate units, the stabilized power supply being external to the instrument proper. The two units are housed in metal cases with carrying handles, and are connected together by means of a flying lead.

EE 05 756 *for further details*

GLASS CRYSTAL UNITS

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

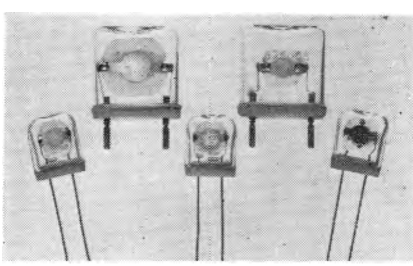
(Illustrated below)

Mullard Ltd has announced some new additions to its range of miniature and sub-miniature glass-encapsulated quartz crystal units. Now an AT-cut glass-encapsulated unit is available for a range of frequencies generally covered in metal holder styles (1·6 to 87MHz).

Compared with metal types, all-glass crystals have an improved ageing performance, a closer initial tolerance and a higher activity.

The miniature size (style L) has a lower frequency limit of 2·3MHz. The sub-miniature types (styles M and N) now cover the range 7·0 to 87MHz in fundamental, 3rd overtone and 5th overtone units. In addition units can be supplied to cover frequencies between 1·6 and 2·3MHz in a longer holder version of the style D, equivalent to the American holder type HC-28/U.

Mullard all-glass crystal units are guaranteed to have an ageing characteristic of not more than +3·5 in



10^6 change in frequency after 90 days at 85°C. In practice this means that in normal usage the change in frequency due to ageing will not exceed about +3 in 10^6 over a period of one year. By mounting the crystal plate in an evacuated, all-glass envelope the mechanical loading on the crystal plate is reduced. This decreases the equivalent series resistance and thus increases the Q-factor.

EE 05 757 *for further details*

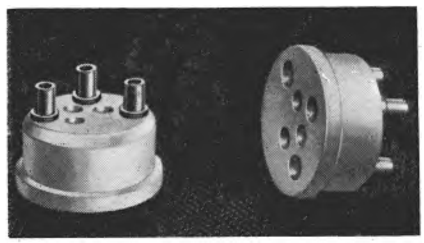
PRESS-FIT TRANSISTOR HOLDER

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustrated below)

Sealectro Ltd has added a transistor holder to its "Press-Fit" range which will accommodate either T0-5 or T0-18 cans having three leads. Designated the RTC-305, the holder has three brass feed-through lugs which accept the leads of the T0-5 can. On a circumference inside that of the lugs are three holes through the Teflon bushing through which leads of the T0-18 holder are fed and wrapped around the lugs.

For positive installation of the RTC-305 at production rates Sealectro offers a special tool designated B-29-X1.



EE 05 758 *for further details*

MINIATURE WELDING TOOLS

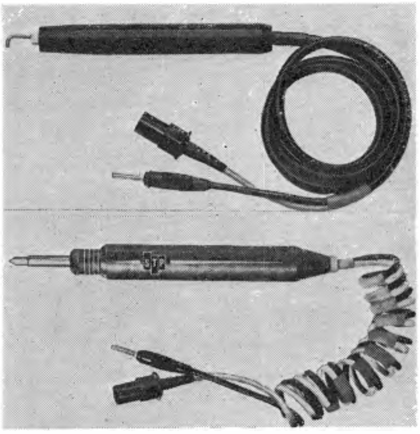
Spembly Technical Products Ltd, Trinity Trading Estate, Sittingbourne, Kent

(Illustrated on page 268)

S.T.P. Ltd announce two new accessories for use with the WPL 4 capacitor discharge power pack. These are the miniature spot welding pencil and the micro spot welder. Both items are small, light and portable and lend themselves particularly to difficult *in situ* welding operations.

The miniature spot welding pencil is suitable for welding fine wires and foil of between 0·0005 and 0·010in thickness and has successfully been used to weld resistance wires of 0·0005in diameter without impairing the resistivity of the wire. Welding current is supplied to the tip when a microswitch, within the body of the pencil, is actuated. The pressure required to actuate the microswitch can be varied by adjusting the length of the tip, enabling an operator to preset the pressure for any specific weld.

The micro spot welder is also pressure sensitive and can be set to operate over a range of welding pressures to suit various materials between 0·0005 and



0·030in thickness. Pressure is set by adjusting a knurled nut on the electrode holder and the correct pressure is indicated on engraved rings on the electrode sleeve.

EE 05 759 *for further details*

SCREWLESS PLUGS AND TERMINALS

Distributed by: Teckna (Sales) Ltd, 47 Whitehall, London, S.W.1

(Illustrated below)

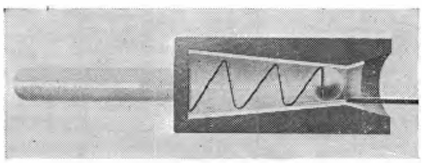
The 'Modena' range of screwless plugs and terminals completely eliminates screw-type connexions, by utilizing a very efficient but simple connexion system.

Each plug and terminal incorporates a spring which exerts pressure on a metal ball. It is therefore only necessary to strip a wire and insert it into the end of the plug. The wire is immediately clamped by the ball and even the hardest wire pull will only result in an increased spring pressure, thus making the contact all the more effective.

Release of the wire can only be effected by applying a counter pressure on the ball, with a pointed instrument.

This fully patented system has many advantages in that there can be no loosening or loss of screws through vibration, wires cannot be damaged in any way, loose contacts are eliminated and no tools are required for mounting.

The largest single advantage, however, is the amount of time which can be saved over conventional connexion methods.



EE 05 760 *for further details*

TRANSISTOR LEAD PREFORMING TOOL

Kingham Electronics Ltd, 17 Briary Wood Lane, Welwyn Heath, Welwyn, Hertfordshire

(Illustrated above right)

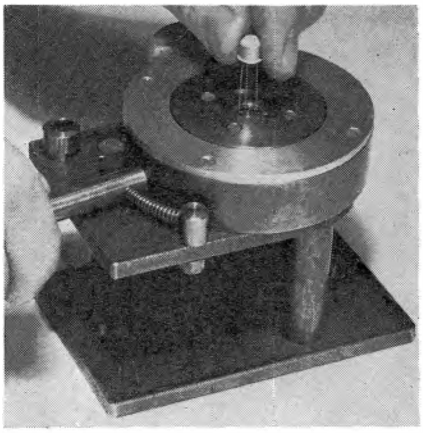
The tool illustrated is the first of a series to be produced, and is for the performing of T0-5 (or other transistors with three leads on a 0·200 bolt circle) transistors prior to their assembly to printed circuit boards.

The transistor leads, which must be straight, are fed through the three holes, as shown in the photograph, until the can is seated on the cover plate. The lever is then pulled and the resultant action joggles and crops the leads to a predetermined length in one smooth operation.

The processed transistor is then removed and snapped into position in the printed circuit board and will remain in place while the board is turned over ready for soldering.

Although this tool is manually operated very little effort is required and it can be used quite easily by female operators.

The loading and removal of the transistor has also been greatly simplified with a view to maintaining a good production turnover.



EE 05 761 *for further details*

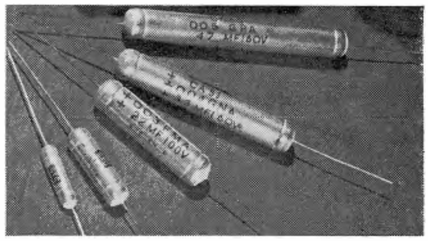
ETCHED TANTALUM FOIL CAPACITORS

Standard Telephones and Cables Ltd, Capacitor Division, Brixham Road, Paignton, Devon

(Illustrated above right)

The Capacitor Division of Standard Telephones and Cables Ltd has announced that its range of etched-tantalum-foil capacitors has received full Qualification Approval to DEF Specification 5134/A-5. The approval is the first to be granted to etched-tantalum-foil capacitors of British manufacture.

Etched-tantalum-foil capacitors are miniature components of very high capacitance, performance and reliability intended for use where miniaturization is an essential requirement. They are similar in design to plain-tantalum-foil capacitors except that the foil is treated with an electro-chemical etching process which considerably increases the foil surface area—as much as seven times—and results in a correspondingly large increase in capacitance-voltage (μFV) product for a given case size. This improvement in efficiency means that for a given μFV product, the cost per μF of an etched-tantalum-foil capacitor is con-



siderably lower than that of an equivalent plain-foil capacitor. Low leakage current and long shelf life are other notable attributes.

Etched tantalum foil capacitors can therefore be used in place of foil-and-paper, metallized-paper, aluminium electrolytic, or, of course, plain tantalum foil capacitors.

STC etched-tantalum-foil capacitors are available in values from 0·68μF to 1500μF with working voltages from 6V to 160V d.c. (85°C max.). Produced in polar and non-polar forms, they meet or exceed the test requirements of MIL-C-3965/2A.

EE 05 762 *for further details*

RECORDING OSCILLOGRAPH

Distributed by: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

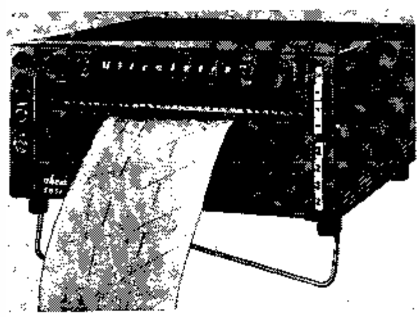
(Illustrated below)

The Instruments Division of Claude Lyons Ltd is marketing the Svenska Diamant 'Ultralette' direct writing mirror galvanometer oscillograph.

Capable of recording up to 30 channels of information simultaneously with frequency response from d.c. to up to 10kHz, the 'Ultralette' 5651 is a high performance instrument incorporating 16 paper drive speeds from 2·5mm/min to 100cm/sec. An accessory high-speed drive unit, and for lengthy records a 25:1 reduction gear, enable the 5651 to cover the very wide overall paper feed range of 1mm/min to 10m/sec.

The 'Ultralette' 5651 incorporates reverse drive for oscillogram rewind, flash timer, pre-set record length and selectable event release. It is wired for full remote control of all 16 standard speeds. The 30cm optical arm and very high quality optics provide excellent sensitivity together with a very fine trace giving superior resolution to competitive oscillographs.

Up to three ten-channel magnet blocks can be fitted to give up to thirty-channel recording capability. Galvanometers available provide natural frequencies



from 20 to 10 000Hz with sensitivities from 0·00058mA/cm to 46·5mA/cm.

The 5651 takes all paper widths from 3in to 12in, the compact design nevertheless providing overall dimensions of only $15\frac{3}{4}$in × $15\frac{3}{4}$in × 7in panel height.

Accessories available include rack mounting adaptors (19in × 7in panel height), writing table with take-up spool, light-proof cover with inspection window, high-speed drive unit, slow-speed reduction gear, paper stabilizing unit, remote control unit and rapid print illuminator (latensifier).

EE 05 763 *for further details*

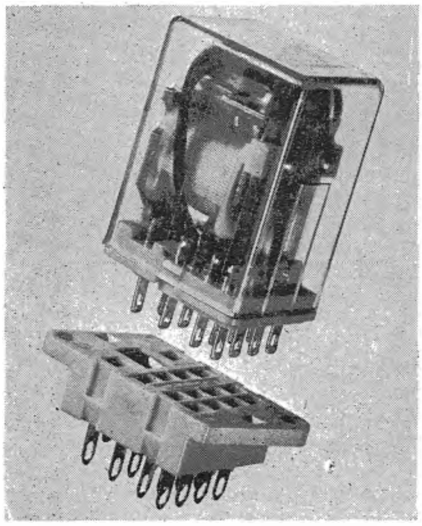
MINIATURE RELAY

Thorn Special Products Ltd, Thorn House, Upper St. Martin's Lane, London, W.C.2

(*Illustrated below*)

Thorn Special Products Ltd is to market a midget relay with contact ratings of 3A 115V a.c. Of robust design and maximum reliability, the RA midget relay uses the standard range of a.c. and d.c. coils with nominal coil voltages of 6 to 115V.

The relay is a four pole change-over unit with fine-silver, gold-flashed contacts. An operating power of 0·8VA, 0·5W is required. Coil rating at 25°C max is 2·5VA or 2W. Operating time of the RA midget relay is 9 to 11msec and release time 3 to 6msec. Insulation resistance is 1500MΩ minimum.



The relay can be mounted and is a plug-in relay with various sockets which can have solder tags or printed circuit terminals. For securing to a socket or printed circuit board the relay has a mounting stud with a 3-42 NC2 thread.

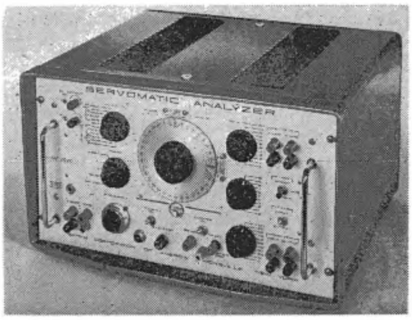
Overall dimensions of the relay including dust-cover are 1·516in high, 1·094in wide and 0·847in deep.

EE 05 764 *for further details*

SERVO ANALYSER

Distributed by: Microwave Systems Ltd, 32 Hunting Gate, Hitchin, Hertfordshire

(*Illustrated above right*)



A Servomatic analyser with new, improved waveform containing less than 1 per cent distortion is now available from Microwave Systems Ltd.

The new analyser provides a single package instrument combining the capabilities of function generation, variable phase reference, internal modulator and precision attenuator. Developed for the design, evaluation and production testing of electronic equipment and systems, the analyser has many new features.

Two main features are the internal modulation which provides a 'bow-tie' image with greater memory during low speed servo analysis and a d.c. offset, which provides for the adjustment of the zero level to a desired bias displacement.

The analyser enables the engineer readily to plot his system's phase and amplitude response to various frequencies in problem applications such as, transfer function of open and closed loops, phase angle measurements, amplitude measurements, transfer function of damping and filter networks, frequency response of portions of systems, computer response and response of magnetic amplifiers.

The instrument combines a multiple function generator with a calibrated phase shifter and precision amplitude control. Sine, square and triangular waveforms are all generated and available in d.c. form or as a modulated carrier. A reference signal is provided which is phaseable with respect to the output signal.

Frequency range is from 0·005 to 1000Hz continuously variable in 5 ranges for all waveforms.

EE 05 765 *for further details*

STABILIZED POWER SUPPLY

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(*Illustrated below*)

A.P.T. Electronic Industries Ltd announce a new model in the TSU 0500 range of stabilized sub-unit power supplies

The new adaption designated 13028/A is a modified version of the standard 0·5A unit, upgraded for a higher current output of 1A at a pre-set voltage of 5V. All silicon transistors are used in place of the germanium transistors and the general construction, appearance and size follows the pattern of the TSU 0500 range.

The 13028/A has been developed to meet a growing demand from industry for a unit of maximum output in the minimum space. It measures an overall $5\frac{3}{8}$in × $3\frac{3}{16}$in × $2\frac{3}{8}$in, the same size as the TSU 0500 range, but its output has been doubled.

Over-current protection is achieved by the use of a circuit which not only ensures that the output voltage will be clamped to zero if it begins to fall, but also limits the output current to a safe value if the load is short-circuited. Provision is also made for the user to obtain further overload protection as required by replacing a link between the transformer secondary winding and the rectifiers with a fuse which can be mounted externally.

EE 05 766 *for further details*

PRE-SET COUNTER

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

(*Illustrated below*)

Venner Electronics Ltd has introduced a five digit pre-set counter for industrial use, type TSA 5545, with a time-base fully variable over five digits.

Essentially a variable time-base frequency meter with facilities for frequency, multi-periods, speed, speed ratio and time measurements, the instrument also provides accurate time delay generation and may be used for batch counting.

The digital time-base can be set to give gating times between 1µsec and 100sec, enabling the counter to display the results of a measurement in any accepted engineering unit such as gallons/minute.

A display storage facility is fitted, enabling the previous result to be retained while the following measurement is made, the display being updated on completion of each measurement; this facility can be switched out of circuit when not needed.

A printer output (1248 code b.c.d.) and a conversion to 19in rack mounting are optional extras.

There are two identical signal inputs (*A* & *B*), with sensitivity of 100mV (max input 250V), impedance 100kΩ min., and frequency range 10Hz to 2MHz. Two identical control inputs, *C* (start) and *D* (stop) accept pulse inputs of 1V to 50V over the frequency range d.c. to 1MHz, with positive or negative slope. Facilities for contact operation are also provided.

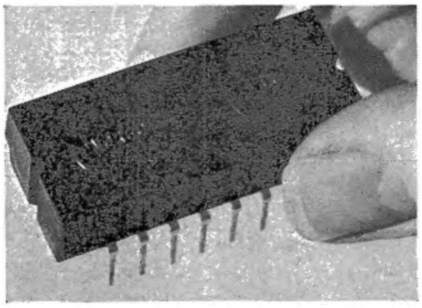
For frequency/ratio uses the TSA 5545 gives a gating time of $n \times (1, 10, 10^2$ or $10^3)$ μsec for frequency, and $n \times (1, 10, 10^2$ or $10^3)$ *B* for ratio, *n* equalling 1 to 99999.

Period measurements (signal input at *A*), gating time $n \times A$, give time units of 1, 10, 10^2, 10^3 μsec, or 1, 10, 10^2, $10^3 \times B$.

Straightforward time measurements are in units of 1, 10, 10^2 or 10^3μsec or input *B*, while delay output is a square-wave starting at input *C* pulse and stopping after $n \times$ the same time units.

End view neon number tubes are used for the display, and a Polaroid screen gives improved contrast. Display time is variable 0·5 to 5sec, or infinite. Accuracy is ±1 count ±crystal stability, which is better than $\pm 1/10^6$. Operating temperature range is from −10°C to +60°C, and power supply 100 to 125 or 200 to 250V, 40 to 60Hz a.c.

EE 05 767 *for further details*



STATIC SWITCHING MODULES

The M.E.L. Equipment Co. Ltd, Manor Royal, Crawley, Sussex

(Illustrated above)

This new type of module, called Norbit 2, is an improved and cheaper version of the static switch module marketed under the name 'Norbits' which M.E.L. has been producing over the past five years. Control systems using the original Norbits instead of relays had the advantage of no moving contacts, no maintenance requirements and a life which was frequently far in excess of that of the equipment in which they were installed. For simple systems, however, the initial cost was higher than for comparable systems employing relays. But now with Norbit 2 the advantages can be obtained at a reduction rather than an increase in cost.

M.E.L. states that it has made a number of studies which confirm that most industrial control systems can be built more cheaply using Norbit 2 modules than with relays. Even with systems which would normally use as few as 15 relays economies can often be effected by using Norbit 2.

The modules are targeted specifically at the industrial control field. Silicon transistors and diodes are used in their construction and the range, in addition to including the much-used NOR function, features an invertor amplifier, timer, input switch filter, power amplifier and power supply unit.

Although Norbit 2 is basically an electronic switch, M.E.L. emphasize that there is no need for users to possess any knowledge of transistor or electronic theory. From the user's point of view the modules can be regarded simply as switches which instead of opening or closing either conduct or do not conduct.

The phsical design of the modules ensures maximum flexibility of mounting and wiring. Measuring 2in × 1in × ½in, they are convenient to handle and yet not too space consuming. From each of the longer sides protrude nine flat pins to which either wrapped wire or soldered connexions can be made. The modules can be secured pins uppermost to a panel or other flat surface by two small screws, or mounted pins downwards on a printed wiring board. Alternatively, sub-assemblies may be built by mounting the modules on pre-drilled metal bars which are also available from M.E.L.

Being encapsulated, Norbit 2 modules are unaffected by dust and can be used with safety in hazardous atmospheres. The permissible working temperature range is quite wide, from −10° to +85°C, and for storage the lower temperature limit extends to −40°C. A feature of the modules is that no special power supplies are required. Operation is from a single rail 24V d.c. supply, which does not have to be stabilized, or via a mains supply unit which is also part of the range.

EE 05 768 *for further details*

ULTRASONIC WIRE BONDER

Vacwell Engineering Co. Ltd, Willow Lane, Mitcham, Surrey

(Illustrated above right)

An ultrasonic wire bonder with magazine facility, for use in the production of integrated circuits and transistors, is announced by Vacwell Engineering Co. Ltd, an EMC Group subsidiary which specializes in plant and test equipment for the semiconductor components industry.

A special feature is the design of the magazine, which allows each workpiece to be moved in two axes and rotated through a full 360° while *in situ* under the bonding head. This eliminates the accepted, but tedious and sometimes costly, handling of the items individually when ultrasonic bonding is normally employed. By this means, completion rates well in excess of 1 600 bonds per hour are readily achieved.

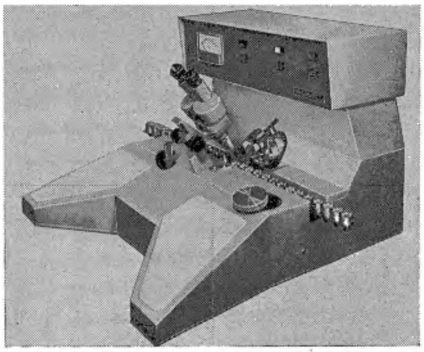
With the new Vacwell bonder, the components to be processed are first placed into simple linear carriers. The loaded carriers are then placed one after another into the machine, and are ejected from the opposite side as the work is completed. This arrangement permits a smooth, uninterrupted flow of assemblies into and out of the bonder at a completion rate which is directly controlled by the operator.

All normal device packages can be dealt with, including dual-in-line, flat packs and TO-5-type multi-lead headers. Substrates of shapes and sizes to customers' specifications can also be accommodated.

Wires of aluminium or gold can be bonded to contacts of the same materials on silicon or silica bases. The wire feed and cut-off system is completely automatic and can handle wires ranging from 10 to 75 microns in diameter. Various forms of bond, including stitch-type, can be produced.

Rotation of the vacuum chuck holding the component is effected through a servo mechanism controlled by a single joystick-type manipulator. A full 360° rotation in the plane of the device is provided. In addition to this rotary movement, the joystick controls movement about the *X* and *Y* axes.

A fixed-magnification stereo microscope is included in the specification, as



are a number of work magazines, whose length and capacity can be specified by the user to suit the component to be processed and the production rate required. The normal capacity of each magazine is 20 or more components.

The bonder is very compact and can be mounted on any work bench or table. While it is intended for use under high-speed production-line conditions, the versatility of the machine can be exploited in a research laboratory or other department concerned with small-scale production.

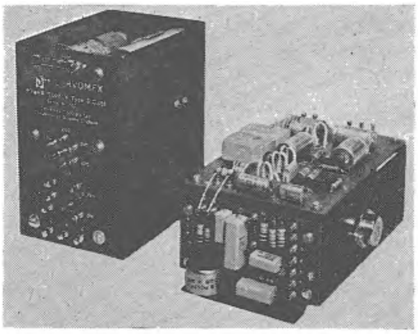
EE 05 769 *for further details*

SERVO AMPLIFIER

Servomex Controls Ltd, Crowborough, Sussex

(Illustrated on page 271)

Servomex Controls Ltd has developed a servo amplifier and associated power unit specifically for use in conjunction with the two-phase servo motor types SM95 and SM135.



These new instruments, SA.186 and power unit DC.151 make the servo motors adaptable for numerous applications in various types of 50 to 60Hz a.c. servo systems.

Two summing resistors are provided and there is a terminal to connect directly with the SM135 tacho-generator output terminal for stabilizing velocity feedback. When used with this servo motor, full torque will result with 4mV into one of the resistors (the other one being earthed) and 2mV will produce full torque if the other resistor is left open-circuit.

The phase shift is about 12° leading at 45Hz and 12° lagging at 65Hz. Thus in 50 to 60Hz supplies the full rated torque will be available.

When connected to the SM95 servo motor, full torque will result with 35mV into either of the summing resistors or with both resistors in parallel full torque is available for 17mV. Typical phase shift is about 75° lag at 45Hz changing to about 110° lag at 65Hz. (The change from the nominal 90° is hardly significant).

The amplifier can be used with 'floating' input. One of the amplifier input terminals is permanently joined to one of the h.t. rails but this does not have to be earthed. The metal work is intended to be earthed and the amplifier and power unit circuits can be 'floated' up to ±50V with respect to the metal work.

Both units are produced in sub-assembly form for inclusion in other equipment. The instruments use silicon semiconductor circuits and can be employed at ambient temperatures up to 70°C.

It should be noted that the amplifier is not suitable for use with any other motor and that no other source of power is recommended.

EE 05 770 *for further details*

PRINTING REPERFORATOR

Creed & Co. Ltd, Hollingbury, Brighton, Sussex

(Illustrated above right)

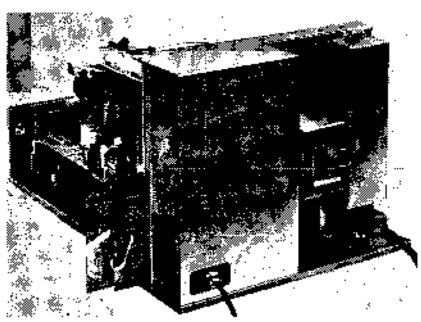
The introduction of a printing reperforator which employs electronics for all functions other than print-out and paper handling has been announced by Creed and Co. Ltd.

The new machine will be known as the Creed EPR electronic printing reperforator and is designed to record telegraph messages and other data both as printed characters and code perforations on standard 5-track $^{11}/_{16}$in wide paper tape. The characters are printed between the feed holes, 8½ feed hole pitches behind the corresponding punched codes.

Fully transistorized, the model EPR incorporates all the facilities of a conventional printing reperforator and is fully compatible with all existing mechanical telegraph equipment.

The use of electronics in the EPR has made the machine more reliable than comparable mechanical equivalents. In all, electronic units based on high reliability silicon-planar transistors have been used to replace 70 per cent of the mechanical parts found in conventional machines. All the mechanical units in conventional machines that require regular skilled maintenance have been eliminated in the EPR, which has no clutch, mechanical selector or motor speed governor. The reduction in moving parts has reduced fault liability and the amount of maintenance attention required. It is estimated that the EPR will require periodic maintenance only once every 1000 hours operation at 50 bauds, or one year, whichever is the earlier.

An additional advantage resulting from the use of electronics is a reduction in the list of essential spares. This has been reduced by 50 per cent through the change in emphasis from mechanical to electronic parts. In the electronics unit there are few special components and the number of component types have been kept to a minimum—nearly all available from any commercial supplier.



Normally the EPR is arranged to operate at speeds of 50 and 75 bauds (66 and 100 words per minute), but other standard speeds are possible. A feature of the EPR is the easy change of speed which can be effected by the repositioning of a simple connecting link in the electronics unit.

The EPR is available either complete with an integral silencing cover or without a cover for tape-relay console mounting. The absence of a speed governor and use of an asynchronous induction driving motor has made it possible to enclose the whole drive unit for quieter operation.

Optional extras available include a priority character recognition facility and manual control of letters/figures shift.

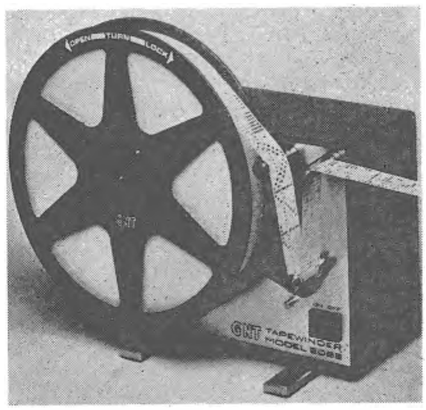
EE 05 771 *for further details*

PAPER TAPE WINDER

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

(Illustrated below)

The GNT model 2082 tape winder is capable of winding paper tape at up to



120 char/sec and has the added advantage that no pull is exercised on the tape whether it be from reader or perforator. Slack tape to the winder controls via a swinging arm the drive through a disk clutch and in the tight tape condition the drive is disengaged and brake pads are applied to prevent over-running of the tape spool.

The tape spool is manufactured in an unbreakable material and is of the 'take-apart' type where by a 45° anti-clockwise rotation of one flange relative to the other parts of the spool, a centre split collar can be removed from the central hub for easy threading of tape or when the reel is full for storage purposes.

EE 05 772 *for further details*

MICROWAVE FREQUENCY DIVIDER

Hewlett Packard Ltd, 224 Bath Road, Slough, Buckinghamshire

(Illustrated on page 272)

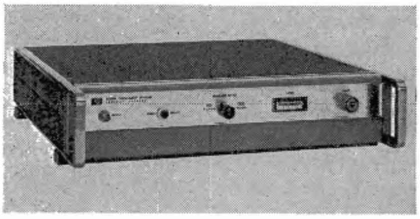
An automatic frequency divider based upon new techniques in frequency division enables high-frequency electronic counters to measure microwave frequencies through X-band (to 12·4GHz). The display is direct and complete, requiring no further calculations. The new instrument (HP model 5260A) operates over all of the most-used portion of the microwave region, 0·3 to 12·4GHz. It accurately divides input frequencies up to 1·2GHz by a factor of 100 or, when the input is within the range of 1·0 to 12·4GHz, by a factor of 1000. The divided-down frequency (1 to

12·4MHz) is supplied to the counter. The divider introduces no error of its own and measurements thus are as accurate as the electronic counter.

Other than selection of the division ratio, the automatic frequency divider requires no adjustments or tuning. Hence, the instrument is well-suited for unattended monitoring of frequency over an extremely broad range.

Read-out on the associated counter is direct. No computations are required except for placement of the decimal point (an optional version of the HP model 5245L 50MHz counter, model MO7-5245L, automatically positions the decimal point when used with the new frequency divider.

The new frequency divider is, in a sense, an automated transfer oscillator but contains elements of both heterodyne convertors and transfer oscillators. An internal oscillator is phase-locked to a submultiple of the input frequency. A



second oscillator is phase-locked to the first one in such a way that exactly one-hundredth or one-thousandth of the input frequency can be derived. No frequency error is introduced and measurements are made with the basic accuracy of the associated counter.

The frequency divider works with input signal amplitudes greater than 100mV and a front panel meter indicates when the signal level is within the proper amplitude range (−7dBm to +10dBm). Input impedance is 50Ω and the v.s.w.r. is relatively constant and less than 1·6:1 up to 10GHz. V.S.W.R. is less than 2:1 up to 12·4GHz.

EE 05 773 *for further details*

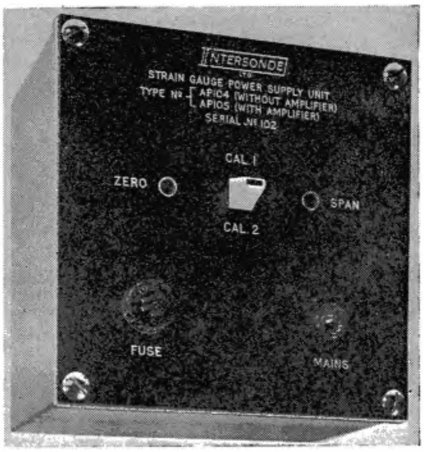
STRAIN GAUGE POWER SUPPLY AND BRIDGE BALANCE UNIT

Intersonde Ltd, The Forum, High Street, Edgware, Middlesex

(Illustrated below)

The new Intersonde type AP104 strain gauge power supply and bridge balance units are available in single and six-channel form and are primarily designed to provide stabilized d.c. excitation of up to 24V for transducers based on resistance strain gauges. The six-channel version features six identical and fully-floating d.c. supplies and both types are suitable for energizing transducers based on strain gauge elements connected in either a half-bridge or full-bridge configuration.

In addition to providing transducer energization the units also embody



controls, in the form of high resolution multi-turn helical potentiometers, for adjusting the span and zero of the transducer. Shunt calibration facilities are also provided and enable a choice of two internal high-stability resistors to be switched across one transducer bridge-arm to produce a known amount of bridge unbalance and hence transducer output.

The units operate from a 200 to 250V, 40 to 60Hz supply and have an output voltage stability of better than 0·1 per cent for a ±10 per cent change in mains input. The temperature coefficient of output voltage is less than 0·003 per cent/°C and the units will produce an output current of up to 100mA over the operating temperature range 0 to 60°C.

EE 05 774 *for further details*

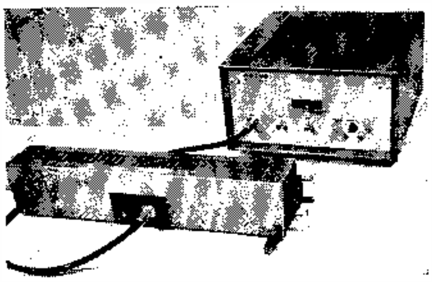
R.F. EXCITED GAS LASERS

System Computors Ltd, Fossway, Newcastle upon Tyne, 6

(Illustrated below)

Two helium neon gas lasers which can be used for a wide variety of experiments and applications are part of a range being manufactured by System Computors Ltd. The lasers, which are r.f. excited, are designated model G1 and model G2.

Model G1 is suitable for demonstrating optical phenomena and for applications such as holography, which require relatively high outputs. It has a 115cm discharge tube with a guaranteed life of three hundred hours. Output of this laser at 6328°A is guaranteed at 1·0mW in single mode and 3·0mW in multi-mode.



The excitation frequency is 30MHz. Hardcoated dielectric mirrors are used and mirrors are available for operation at other wavelengths or for single ended output.

Model G2 operates at the same wavelengths as model G1 but in standard form it has the added facility that the laser beam can be modulated, which makes it especially suitable for communications applications. To achieve this there is an amplitude modulation stage in the power unit. Output power is 0·4mW in single mode and 1·0mW in multi-mode. It has a 50cm discharge tube and the r.f. excitation frequency is 27·1MHz. Accessories available include a power output meter and an amplifier loudspeaker unit.

EE 05 775 *for further details*

RESISTANCE BOX

Shirehall Electronics Ltd, Station Yard, Borough Green, Sevenoaks, Kent

(Illustrated above right)

The Shirehall *R*-box provides quick selection of resistance, in preferred values, from 1Ω to 6·8MΩ. With only two controls, one for the tens multiple, it is very easy to use and read the value selected.

From 1Ω to 6·8Ω, the internal resistors are wirewound and rated at 3W; from 10Ω to 6·8MΩ they are 0·5W, high stability carbon film. Tolerance throughout is 5 per cent.

The *R*-box is intended as a general purpose aid for design and development engineers and will find wide application in selection of optimum values, fault-finding by substitution and production testing of sub-assemblies and complete equipment.

It is fitted with an 'earthing' terminal, connected to the metal case for screening purposes, and supplies complete with flexible leads and hermaphrodite connectors.

EE 05 776 *for further details*

INSTRUMENTATION TAPE HEADS

Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

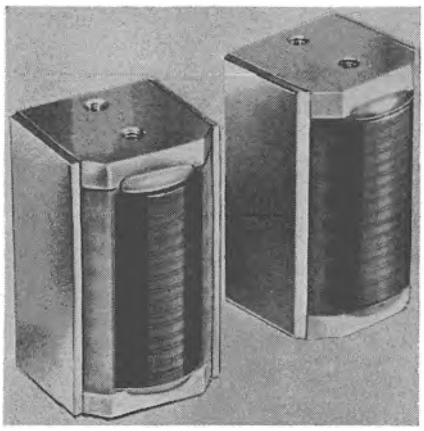
(Illustrated on page 273)

A new range of high-quality record and replay high-density ferrite magnetic heads for multi-track analogue instrumentation recorders is now available from Mullard Ltd.

The heads are high-density ferrite types in which glass/ceramic constructional techniques have been used to produce a tape contact area which has a virtually unlimited life. The resulting consistent performance completely eliminates the frequent circuit compensatory adjustments which are necessary when other types of head are used. Other interesting features are precise gap alignment and improved h.f. performance, the latter being due to the low inherent core losses of the ferrite material. Low core losses also ensure excellent performance at low signal levels and establish the bias required for optimum performance at a much lower level than usual.

Conforming to the IRIG recommendation for direct and f.m. operation, these Mullard heads are compatible with those used in all existing IRIG system recorder equipments.

Pairs of the 7-track heads can be doubled up on a suitable mounting base



to provide a 14-track 1in interlaced system to IRIG standards, while the 3 and 4-track heads can similarly be arranged in a 7-track 0·5in IRIG interlaced system.

Details of the six heads in the new range are given below:

ER7300 Record 1in tape
7 data tracks + 1 annotation track
ER7301 Replay
ER7302 Record 0·5in tape
3 data tracks + 1 annotation track
ER7303 Replay
ER7304 Record 0·5in tape
4 data tracks
ER7305 Replay

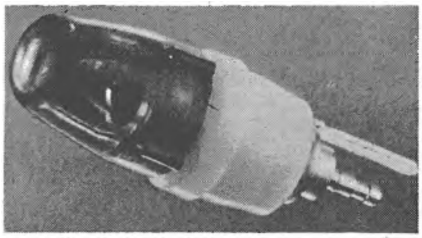
EE 05 777 *for further details*

MINIATURE INDICATOR LAMP

Oxley Developments Co. Ltd. Priory Park, Ulverston, Lancashire

(Illustrated above right)

A miniature indicator lamp has recently been introduced by Oxley Developments Co. Ltd. The design is



based on the patented "Barb" principle which ensures firm anchorage in chassis or panels. Assembly consists of pressing the insulating bush into an ordinary 0·281in drilled or punched hole.

The lens caps for these components are available in six colours: red, amber, yellow, green, blue, clear.

The Oxley 'Barb' indicator has been designed to incorporate and utilize fully the advantages of a low voltage, low consumption, subminiature incandescent lamp—suitable for a wide variety of applications such as compact programming displays, medical equipment, airborne electronic equipment, etc.

EE 05 778 *for further details*

MINIATURE RELAY

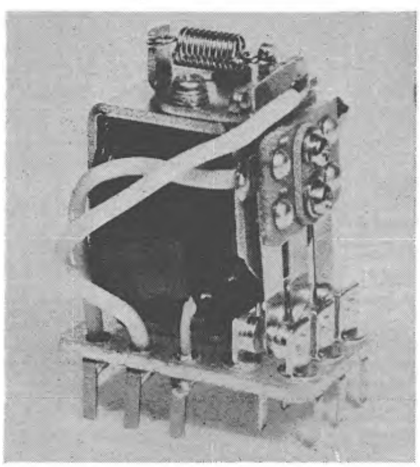
Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustrated below)

Now available from Keyswitch Relays Ltd is a miniature relay (type MK3—0) designed for direct printed circuit insertion, with all terminations including those from the coil becoming part of the complete terminal board.

Characteristics of this relay are similar to the basic MK3 (3-pole, double throw) relay designed for applications in control and production systems, conventional power stations, chemical industries, radio and television transmission and general electrical and mechanical engineering.

Switching current of the MK3 -0 is 7·5A per contact, with maximum contact voltage 440V a.c. or 250V d.c. Coil consumption is not more than 2·5VA a.c. or 1W d.c. The relay has a life in excess of 5 000 000 operations and a fast operating time of 15msec (a.c.) or



25msec (d.c.). Release times are 20msec (a.c.) and 30msec (d.c.), and insulation 1 000V a.c.

The Keyswitch MK1 and MK2 miniature relays are also available for printed circuit use.

EE 05 779 *for further details*

TELEMETERING SYSTEM

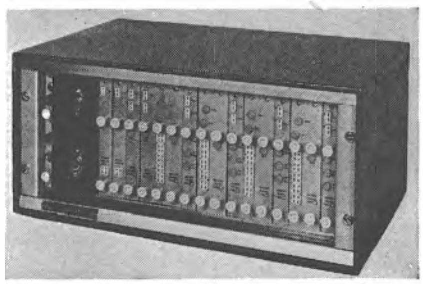
Telectron Ltd, 16 Herbert Street, Dublin, 2, Ireland

(Illustrated below)

The AC-30 telemetering system is a completely solid state, tone frequency multiplexing system, providing up to 30 voice frequency channels in the range 420Hz to 3 900Hz suitable for the continuous transmission of analogue, on/off supervisory and remote control signals.

Telephone lines, radio links, carrier channels or a pair of wires may be used as bearer circuits.

Measurement data is transmitted, over an analogue channel, with extreme accuracy, the error being 1 per cent or less. A pulsed frequency system is used whereby the level of the quantity to be measured determines the number of pulses



per second, of the particular channel frequency, transmitted across the bearer circuit while the d.c. output at the receive end is directly proportional to the number of pulses per second received.

Indication and control functions are handled by on/off channels which consist of highly selective tone generators and receivers, one channel being required for each function. Application of an earth to a tone generator causes the reed relay in the output of the tone receiver to be energized.

The system is designed on the modular building block principle and can be expanded from a minimum capacity of one channel by the addition of further modules.

Particular applications are the remote control and supervision of waterworks, pumping stations, electricity transformer stations, meteorological stations, radio and television stations and microwave link repeater stations, sewage plants, gas and oil pipe lines etc.

This equipment is marketed in the U.K. by Associated Electrical Industries Int. Ltd, Telecommunications Division, Woolwich.

EE 05 780 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All evening meetings will be held at Savoy Place, commencing at 5.30 p.m. (tea at 5 p.m.), unless otherwise stated.

Date: 4 April.
Colloquium: On-line Scheduling for Production Purposes.
(Joint meeting with the Automatic Control Group of the I.Mech.E. Time to be announced in IEE News).

Date: 4 April.
Discussion: Discharge Detection: Integrated or Resolved Response?
Opened by: G. Mole.

Date: 5 April.
Discussion: Distribution System Control.
Opened by: R. H. Rockliffe and J. R. Hill.

Date: 7 April.
Lecture: Digital Computer Programme for Design Synthesis of Large Squirrel-Cage Induction Motors.
By: B. J. Chalmers and B. J. Bennington.

Date: 7 April.
Lecture: Development of Fluid Damped Electrical Measuring Instruments.
By: S. W. Eckett.

Date: 10 April.
Lecture: Electronics in Aircraft.
By: R. Cockburn.
(Joint meeting with the Institution of Electrical and Electronic Technician Engineers, at 6 p.m., tea at 5.30 p.m.).

Date: 10-14 April.
Easter School on the P.A.L. Television System.
(At the University of Nottingham).

Date: 11-14 April.
Advances in Computer Control.
(Second U.K.A.C. Control Convention at Bristol).

Date: 11 April.
Lecture: Transient Torques in Induction Motors due to switching off the Supply.
By: W. S. Wood, F. P. Flynn and A. Shanmugasundaram.

Date: 11 April.
Lecture: Influence of Transient Switching Currents and Fluxes on the Torque Developed by a Squirrel-Cage Induction Motor.
By: N. C. Enslin, W. M. Kaplin and J. L. Davis.

Date: 11 April.
Lecture: Digital Computation of Induction Motor Transient Torque Patterns.
By: R. D. Slater, W. S. Wood, F. P. Flynn and R. Simpson.

Date: 11 April.
Lecture: Transient Performance of the Induction Motor.
By: I. R. Smith and S. Sriharan.

Date: 11 April.
Lecture: Note on Induction-Machine Performance with a Variable-Frequency Supply.
By: P. J. Lawrenson and J. M. Stephenson.

Date: 11 April.
Lecture: Time-Domain Analysis of the Invertor-Fed Induction Motor.
By: E. E. Ward, A. Kazi and R. Farkas.

Date: 11 April.
Lecture: Induction-Motor Reswitching Transients.
By: I. R. Smith and S. Sriharan.

Date: 12 April.

Lecture: The Future Use of Solid-State Devices in the Microwave Field.
By: J. E. Carroll.
(At 6 p.m., tea at 5.30 p.m.).

Date: 14 April.
Lecture: A Survey of Semiconductor Materials.
By: S. E. Bradshaw.

Date: 20 April.
Technical Visit to the Clarendon Laboratory, Oxford.
(Further details to be announced in IEE News).

Date: 21 April.
Discussion: The Dynamic Measurement of Stress Using Strain Gauges.
(Joint meeting with the Society of Instrument Technology).
Opened by: D. J. Male.

Date: 24 April.
Lecture: Thirteenth Graham Clark.
By: The Rt. Hon. the Lord Bowden.
(At the I.C.E., 1-7 Great George Street, London, S.W.1.).
(At 5.30 p.m., tea at 5 p.m. Further details to be announced).

Date: 24 April.
Discussion: Heterojunctions.
Opened by: G. T. Wright and T. L. Tamsley.

Date: 25 April.
Discussion: In-Process Gauging.
(Joint meeting with the Automatic Control Group of the I.Mech.E.).
Opened by: R. Sharp.
(At the I.Mech.E., 1 Birdcage Walk, London, S.W.1, at 6 p.m., tea at 5.30 p.m. Further details available from the Group's Department, I.Mech.E.).

Date: 25 April.
Discussion: The Processing of Biological Signals.
(Joint meeting with the I.E.R.E. Medical and Biological Electronics Group at 10 a.m.).

Date: 26 April.
Discussion: The Reduction of Dynamical System Complexity.
(Joint meeting with the Automatic Control Group of the I.Mech.E.).
Opened by: J. H. Anderson, A. D. McCann and D. Mitra.

Date: 27 April.
Lecture: Recent Studies in Surface Physics (58th Kelvin Lecture).
By: F. P. Bowden.

Date: 28 April.
Lecture: Attenuation of 8·6mm-wavelength Radiation in Rain.
By: T. W. Harrold.

Date: 28 April.
Lecture: The Prediction of Attenuation by Rainfall on Line-of-Sight Microwave Paths.
By: B. J. Easterbrook and D. Turner.

Date: 28 April.
Technical Visit to Technical Training Establishments of British European Airways.
(Further details to be announced in IEE News).

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Date: 12 April.
Time: 6 p.m.
Held at: Institution of Electronic and Radio Engineers, 8-9 Bedford Square, London, W.C.1.
Lecture: Circuit Design using Digital Computers.
By: E. Wolfendale.

Date: 25 April.
Time: 10 a.m.
Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Symposium: Processing of Biological Signals.
(Tickets necessary. Joint I.E.E.-I.E.R.E. Medical and Biological Electronics Group).

Date: 26 April.
Time: 2.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Symposium: Ultrasonic Methods in Non-Destructive Testing.
(With Electro-Acoustics Group).

Thames Valley Section

Date: 11 April.
Time: 5 p.m.
Held at: The Lecture Theatre, School of Electronic Engineering, R.E.M.E., Arborfield, Berkshire.
Lecture: Infra-Red Detection.
By: S. D. Smith.

Date: 26 April.
Time: 7.30 p.m.
Held at: Rutherford High Energy Laboratory, Chilton, Didcot, Berks.
Lecture: Instrumentation for High Energy Physics Experiments.
By: P. Wilde.

South Western Section

Date: 26 April.
Time: 7 p.m.
Held at: University of Bristol, University Walk, Clifton, Bristol 8.
Meeting: Super Conductivity.
Opened by: J. G. G. Milne.
(Joint with the Institute of Physics and the Physical Society (S.W. Section).

West Midland Section

Date: 10 April.
Time: 7.15 p.m.
Held at: Lanchester College of Technology, Priory Street, Coventry.
Lecture: Parametic Amplifiers.
By: D. P. Howson.

Date: 26 April.
Time: 7.15 p.m.
Held at: Wolverhampton College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Masers and Lasers.
By: D. C. Laine.

Scottish Section

Date: 19 April.
Time: 7 p.m.
Held at: Department of Natural Philosophy, Edinburgh University.
Lecture: Engineering Development in BBC External Services.
By: K. R. Sturley.

Date: 20 April.
Time: 7 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow C.2.
Lecture: Engineering Development in BBC External Services.
By: K. R. Sturley.

East Anglian Section

Date: 26 April.
Time: 7 p.m.
Held at: Harlow Technical College, The High, Harlow, Essex.
Lecture: Waveguides.
By: A. L. Cullen.

Yorkshire Section

Date: 13 April.
Time: 7 p.m.
Held at: Leeds University, Department of Electrical Engineering.
Film evening: 'Thin Film Microcircuits', 'Particles Count', 'Electrons in Harness', 'Conquest of the Atom', 'Transistors'.

Merseyside Section

Date: 12 April.
Time: 7 p.m.
Held at: Liverpool College of Technology, Byrom Street, Liverpool.
Lecture: Electronic Melodic Instruments.
By: K. A. Macfadyen.

North Eastern Section

Date: 12 April.
Time: 6 p.m.
Held at: The Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle.
Lecture: Radio Astronomy.
By: I. Morison.

South Midland Section

Date: 18 April.
Time: 7 p.m.
Held at: The BBC Club, High Street, Evesham.
Meeting: Central Electricity Generating Board Communications System.
Opened by: K. S. Hooper.
(Joint with the Worcestershire Sub-Centre of the I.E.E.).

THE ROYAL TELEVISION SOCIETY

Date: 10-14 April.
Easter School on the P.A.L. Television System.
Held at: The University of Nottingham.

Date: 21 April.
Lecture: The Strange Journey from Retina to Brain (The Fleming Memorial Lecture).
By: R. W. G. Hunt.
Held at: The Royal Institution, Albemarle Street, London, W.1.
(Tickets will be necessary).

Date: 27 April.
Meeting: Annual General Meeting.

Résumés des Principaux Articles

Théories simplifiées d'information et de transmission de données par A. E. Cawkell

Résumé de l'article aux pages 212 à 218

La théorie de l'information est expliquée avec un minimum de données mathématiques et quelques "racourcis". Il en résulte un article d'une lecture aisée à l'intention des profanes dans ce domaine. Après avoir étudié le contenu des messages d'information, l'article examine la vitesse de transmission des signaux—considérés comme éléments discrets et continus—à travers un canal de communication bruyant. Le concept d'entropie est présenté séparément. L'article traite ensuite du codage, des échanges de largeur de bande, de la puissance, du temps, des méthodes de correction et de dépistage d'erreurs, du minutage et de la synchronisation et, enfin, des systèmes de modulation. Les exemples d'application de la théorie sont limités aux transmissions de données par lignes téléphoniques en raison de l'intérêt porté actuellement à l'emploi des réseaux publics à cet effet. Plusieurs systèmes modernes sont décrits en détail.

Un filtre actif simple avec commande indépendante des emplacements de pôle et de zéro par W. Farrer

Résumé de l'article aux pages 219 à 222

Cet article décrit un dispositif de circuit actif utilisant un réseau en parallèle T modifié, un amplificateur avec gain positif unitaire et deux potentiomètres permettant des caractéristiques de filtre passe-bas ou passe-haut.

La fonction de transfert dans chacun des cas consiste en deux pôles et deux zéros conjugés. La variation de l'un des potentiomètres fixe l'emplacement du zéro sur l'axe de fréquence tandis que l'autre potentiomètre règle la position des pôles de manière à ce qu'on puisse obtenir une réponse linéaire maximale dans la caractéristique de passe-bande.

Deux circuits d'entraînement de borne pour vibrateur mécanique entraîné piézoélectriquement par T. J. Stevenson

Résume de l'article aux pages 222 à 223

La réalisation d'un circuit pouvant entraîner un vibrateur mécanique au moyen d'un seul transducteur piézoélectrique à résonance est compliquée par la capacité relativement importante du transducteur. Une réalisation appropriée est indiquée en considérant le courant dans le transducteur pour un entraînement de tension d'ondes carrées et des circuits sont décrits qui fonctionnent en changeant la tension appliquée entre deux niveaux lorsque le courant inverse les signaux.

Etant donné que les éléments inductifs tendraient à résoné en fonction de la capacité du transducteur, d'autres méthodes sont nécessaires si on veut obtenir l'entraînement d'une tension supérieure à la tension d'alimentation. L'article décrit un tripleur de tension approprié.

Élaboration d'un système radio VHF et MF pour services d'autobus avec contrôle d'horaires par T. K. Tawfig

Résumé de l'article aux pages 224 à 227

Dans le système décrit, une méthode de communication à deux voies entre une station de base et des autobus est employée pour fournir des informations sur la position exacte de chaque autobus en service. Le temps et l'emplacement de chaque autobus est comparé par calculatrice avec un tableau-horaire et la position du moment est indiquée sur une carte au centre de contrôle.

Matériel électronique pour véhicules à moteur par G. N. Patchett

Résumé de l'article aux pages 228 à 232

Le matériel électronique n'est presque pas utilisé dans les véhicules à moteur. Les tubes sont d'un emploi difficile car la haute tension anodique requise ne s'obtient pas facilement. Par contre, les semi-conducteurs s'adaptent bien aux instruments des véhicules à moteur. Les divers dispositifs décrits dans cet article sont entièrement constitués de corps solides et conçus pour assurer une plus grande sécurité d'emploi au véhicule. Trois de ces dispositifs sont destinés à assurer le fonctionnement réglementaire des feux, à savoir, les deux feux de frein, les quatre feux de côté et les feux de côté aux heures d'éclairage. Un autre permet le contrôle des freins et un autre encore de réduire le risque de dépasser accidentellement la limite de vitesse. Enfin, un dernier dispositif permet d'améliorer la vision dans des conditions de pluie intermittente.

Appareil de codage numérique pour l'enregistrement des mesures d'intensité de champ par D. G. Beadle et D. E. Susans

Résumé de l'article aux pages 233 à 239

Cet article décrit un matériel qui a été réalisé pour traduire la sortie de tension continue d'un ou de plusieurs récepteurs de mesure d'intensité de champ sous une forme numérique. Les sorties des récepteurs sont échantillonnées à intervalles réguliers et les valeurs sont enregistrées au moyen d'un code binaire à 5 éléments sur bande de papier pouvant être, par la suite, injecté à une calculatrice aux fins d'analyse.

Les résultats obtenus témoignent d'une concordance satisfaisante avec d'autres méthodes d'analyse.

Un analyseur pour l'électromyographie humaine par P. Fitch

Résumé de l'article aux pages 240 à 243

L'auteur décrit un instrument pour l'analyse automatique de signaux électromyographiques. Prévu pour les diagnostics cliniques, il produit deux trains d'impulsions en série représentant respectivement la formation de crêtes dans les formes d'onde et leur amplitude. La méthode d'analyse consiste à traiter toutes les variations de la valeur efficace de tension sans référence à sa valeur absolue.

La redondance dans les systèmes numériques par F. A. Inskip

Résumé de l'article aux pages 244 à 249

Cet article décrit diverses méthodes pour l'emploi de la redondance dans les systèmes numériques. Des valeurs numériques de fiabilité sont indiquées pour un petit système d'un ordre de grandeur caractéristique des petits systèmes de traitement des données. Deux indices de fiabilité sont fournis, l'un utilisant les valeurs optima de fiabilité des composants pouvant être obtenus actuellement et l'autre utilisant un ordre de valeur inférieur.

Convertisseur pour enregistrement automatique de position mecaniques par U. Kracht

Résumé de l'article aux pages 250 à 251

Un circuit multivibrateur monostable est utilisé pour convertir une valeur de résistance en une impulsion rectangulaire dont la longueur est proportionnelle à la valeur de cette résistance. Le curseur d'un potentiomètre lié au mouvement mécanique au travers d'un système d'engrenages adéquat, agit comme convertisseur de position mécanique-temps. L'impulsion rectangulaire est synchronisée par un trigger extérieur déterminant le moment auquel la position mécanique est mesurée. Cette impulsion est ensuite convertie en un train d'impulsions utilisables pour une lecture visuelle d'un compteur ou envoyé vers un système d'acquisition des données. Le nombre d'impulsions est donc proportionnel à la position mécanique.

La propriété de ce même circuit multivibrateur comme convertisseur tension-temps est brièvement decrite.

Générateurs d'impulsions doubles à tension constante pour les applications électrophysiologiques par A. F. Lewis

Résumé de l'article aux pages 252 à 258

L'auteur analyse les caractéristuqes de réalisation des générateurs d'impulsions doubles utilisés dans l'électrophysiologie. Il souligne, en particulier, la souplesse d'emploi de ces appareils qui leur permet une gamme étendue d'applications. Un circuit constitué de corps solides d'une conception assez simple fournit des impulsions rectilignes appariées et à tension constante, d'amplitude et de durée indépendantes. L'impédance de sortie est très basse (5Ω) et les circuits de minutage sont stabilisés de manière adéquate contre les variations de température et de tension d'alimentation; l'appareil convient donc pour le fonctionnement sur batterie. Les temps de montée et de chute sont de 5 et 10 µsec respectivement et l'ensemble de l'appareil est suffisamment robuste du point de vue électrique pour résister à l'usage prolongé dans les classes d'étudiants.

Zusammenfassung der wichtigsten Beiträge

Vereinfachte Informationstheorie und Datenübertragung von A. E. Cawkell

Zusammenfassung des Beitrages auf Seite 212-218

Die Informationstheorie wird so weit wie möglich ohne Mathematik und mit einigen Kürzungen erklärt, um denjenigen, die sich nicht sehr intensiv mit dem Gebiet befassen wollen, einen lesenswerten Beitrag zu bieten. Einer Besprechung des Informationsgehalts von Nachrichten folgt die Untersuchung der Übertragungsgeschwindigkeit von Signalen, die als diskrete und kontinuierliche Elemente betrachtet werden, in einem verrauschten Nachrichtenkanal. Das Entropiekonzept wird getrennt behandelt. Weitere Abschnitte des Beitrags befassen sich mit Codieren, der Beziehung zwischen Bandbreite, Leistung und Zeit, mit Methoden für Fehlerkorrektur und -erkennung, Taktsteuerung und Gleichlauf, sowie Modulationssystemen. Beispiele der angewendeten Theorie werden auf Datenübertragung über Fernsprechleitungen beschränkt, da zur Zeit Benutzung des öffentlichen Fernsprechnetzes für diese Zwecke im Vordergrund des Interesses steht. Verschiedene moderne Systeme werden eingehend beschrieben.

Ein einfaches aktives Filter mit unabhängiger Kontrolle über die Pol- und Nullagen von W. Farrer

Zusammenfassung des Beitrages auf Seite 219-222

Der Beitrag beschreibt eine aktive Schaltungsanordnung mit einem abgewandelten Doppel-T-Netzwerk, einem Verstärker mit positiver Verstärkung von ein und zwei Potentiometern, so dass man entweder Tiefpass- oder Hochpass-Filtereigenschaften einstellen kann. Die Übertragungsfunktion besteht in jedem Fall aus zwei Polen und zwei konjugierten Nullen. Durch Änderung des einen Potentiometers wird die Lage von Null auf der Frequenzachse eingestellt, während das andere Potentiometer die Lage der Pole so einregelt, dass im Durchlassbereich ein maximal linearer Frequenzgang erreicht wird.

Zweipol-Treiberschaltung für piezoelektrisch getriebene mechanische Vibratoren von T. J. Stevenson

Zusammenfassung des Beitrages auf Seite 222 223

Der Entwurf einer Treiberschaltung für einen mechanischen Vibrator mit nur einem piezoelektrischen Wandler bei Resonanz wird durch die verhältnismässig hohe Kapazität des Wandlers kompliziert. Betrachtung des durch eine Rechteckwellenspannung im Wandler hervorgerufenen Stroms führt zu einem geeigneten Entwurf; Schaltungen werden beschrieben, die durch Änderung der anglegten Spannung zwischen zwei Pegeln arbeiten, wenn der Strom seine Polarität umkehrt.

Da induktive Elemente dazu neigen würden, mit der Kapazität des Wandlers in Resonanz zu schwingen, muss man andere Methoden anwenden, wenn höhere Spannungen als die Speisespannung zum Treiben erforderlich sind. Ein geeigneter Spannungsverdreifacher wird beschrieben.

Planung eines UKW-FM-Funksystems für Omnibusdienste mit Fahrplankontrolle von T. K. Tawfig

Zusammenfassung des Beitrages auf Seite 224-227

Ein Zweiweg-Fernmeldesystem zwischen einer ortsfesten Station und Omnibussen gibt Information über die genaue Position jedes im Dienst befindlichen Omnibusses. Zeit und Örtlichkeit jedes Omnibusses werden im Datenverarbeiter mit einem gespeicherten Fahrplan verglichen und die derzeitige Position auf einer Karte im Kontrollraum angezeigt.

Elektronische Ausrüstungen für Kraftfahrzeuge von G. N. Patchett

Zusammenfassung des Beitrages auf Seite 228-232

Elektronische Ausrüstungen fanden bisher in Kraftfahrzeugen nicht viel Anwendung. Die Verwendung von Röhren ist wegen der erforderlichen hohen Anodenspannugen, die nicht so einfach zur Verfügung stehen, schwierig; degegen lassen sich Halbleiter gut den Stromversorgungen von Kraftfahrzeugen anpassen. In diesem Beitrag werden verschiedene, durchweg in Festkörpertechnik ausgeführte Einrichtungen beschrieben, die die Kraftfahrzeuge betriebssicherer machen. Drei der Einrichtungen garantieren, dass die Lampen den Verordnungen entsprechend arbeiten: beide Bremslichter, alle vier Seitenlichter, sowie Seitenlichter bei eintretender Dunkelheit. Eine andere Einrichtung überwacht, dass die Bremsen in Ordnung sind, eine weitere verringert die Gefahr der zufälligen Überschreitung der zulässigen Höchstgeschwindigkeit. Die letzte Einrichtung gibt bei unterbrochenem Regenwetter bessere Sicht.

Digital-Kodiergerät zur Aufzeichnung von Feldstärkemessungen von D. G. Beadle und D. E. Susans

Zusammenfassung des Beitrages auf Seite 233-239

Der Beitrag beschreibt ein Gerät, das für die Umsetzung des Gleichstromausgangs eines oder mehrerer Feldstärkemessempfänger in Digitalform entwickelt wurde. Die Ausgänge der Empfänger werden in regelmässigen Intervallen abgetastet und die Werte mittels eines 5-Kanal-Binärcodes auf einen Lochstreifen übertragen, der die Computer-Eingabe für die Analyse darstellt.

Die genannten Ergebnisse stimmen gut mit anderen Analysenmethoden überein.

Ein Analysator für Anwendung in der menschlichen Elektromyografie von P. Fitch

Zusammenfassung des Beitrages auf Seite 240-243

Ein Gerät für die automatische Analyse elektromyografischer Signale wird beschrieben. Das für die klinische Diagnose bestimmte Instrument erzeugt zwei Impulsreihen, die das Auftreten von Wellenformspitzen bzw. ihre Amplitude darstellen. In der Analyse werden alle Abweichungen der Signalspannung ohne Berücksichtigung ihrer absoluten Werte verarbeitet.

Redundanz in Digitalsystemen von F. A. Inskip

Zusammenfassung des Beitrages auf Seite 244-249

Verschiedene Methoden zur Anwendung von Redundanz in Digitalsystemen werden beschrieben. Für kleine Systeme, die grössenordnungsmässig typischen kleinen Datenverarbeitungssystemen entsprechen, werden numerische Zuverlässigkeitswerte gegeben. Dabei handelt es sich um zwei Zuverlässigkeitswerte, von denen einer auf den zur Zeit erzielbaren besten Bauelement-Zuverlässigkeitswerten beruht, der andere eine Grössenordnung niedriger liegt.

Ein Umsetzer für automatische Registrierung einer mechanischen Position von U. Kracht

Zusammenfassung des Beitrages auf Seite 250-251

Eine einfache monostabile Multivibratorschaltung ist verwendet als Umsetzer eines Widerstandwertes in einen Impuls, dessen Länge proportional zum Widerstandswert ist. Der Schleiferkontakt eines Potentiometers ist über ein geeignetes Getriebe verbunden mit der mechanischen Bewegung. Dadurch entsteht ein Umsetzer für mechanische Position zu Zeit. Der Impuls wird ausgelöst durch einen "Trigger-Impuls" welcher den genauen Zeitpunkt der Messung bestimmt und danach umgewandelt in eine Reihe von kürzeren Impulsen deren Anzahl proportional zur mechanischen Position ist und daher geeignet für eine Anzeige durch Zähler bzw. zur Verbindung mit Datenverarbeitungsanlage.

Die Eigenschaften der selben Multivibratorschaltung als Umsetzer "Spannung in Zeit" ist auch kurz erläutert.

Konstantspannung-Doppelimpulsgeber für die Elektrophysiologie von A. F. Lewis

Zusammenfassung des Beitrages auf Seite 252-258

Der Entwurf eines Doppelimpulsgebers für elektrophysiologische Zwecke mit betonter Vielseitigkeit, der den Anforderungen eines breiten Anwendungsspektrums genügt, wird besprochen. Eine beschriebene, verhältnismässig einfache Festkörperschaltung erzeugt Konstantspannungs-Rechteckimpulspaare unabhängiger Amplitude und Dauer. Die Ausgangsimpedanz ist sehr niedrig (5Ω), und die Zeitschaltungen sind ausreichend gegen Temperatur- und Speisespannungsschwankungen stabilisiert, so dass sich das Gerät für Batteriebetrieb eignet. Anstiegs- und Abfallzeiten sind 5 bzw. 10 μs, und das Instrument ist elektrisch robust genug, um langzeitigen Missbrauch durch Studenten auszuhalten.