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Commentary

ON the afternoon of Monday, November 2, 1936, the first regular television service in the world was inaugurated by the British Broadcasting Corporation and now, twenty-one years after, is perhaps the occasion to look back and see how far we have journeyed in this comparatively short span.

The launching was a short and unpretentious affair consisting of a few formal speeches followed by two variety turns and a news film; the programme was transmitted from Alexandra Palace and was seen by a very few people, since there were only about 300 television receivers in private hands. It was a modest beginning but the impact it made was like a clap of thunder and few of those early pioneers could have foreseen the enormous growth of television that has taken place in the intervening years or the profound change it would make in our social habits.

The official opening in 1936 was the climax of nearly two years of feverish activity following the publication in 1935 of the Report of the Selsdon Committee which had been set up to study the claims of the rival systems developed by Baird and Marconi-E.M.I.

The former system was complicated both in structure and in use. It employed the clumsy Nipkow disk, while for studio work there were the indirect film method which scanned the film after a delay of about one minute required for the development of the film, and a method of direct scanning employing a number of photo-electric cells. The Baird system was based on 240 lines and 25 frames a second with sequential scanning.

On the other hand, the Marconi-E.M.I. system depended on the newly developed electron camera tube which they named the "Emitron". It possessed greater flexibility in operation and proved successful for outside transmissions as well as for studio work and the Marconi-E.M.I. system promised a higher definition with 405 lines with interlaced scanning.

The Television Advisory Committee, set up by the Government, with true British genius advised the BBC to use both systems and thus it was that the opening ceremony was performed twice, once on the Baird system and once on the Marconi-E.M.I. system and, thereafter, until the Baird system was finally discarded in February 1937, the two systems were used alternately, a week at a time.

In the years that followed, the problems encountered by the BBC with this new science were many, but there was a growing enthusiasm in the country and when the outbreak of war in September 1939 caused the closing down of the television service, it was estimated that the number of television receivers had grown to 20 000, all of which were served by the single transmitter at Alexandra Palace.

In all, the television service of the BBC was off the air for nearly seven years and, during that time, its transmitters were taken over by the Royal Air Force and its staff dispersed to other work. But the war years were not entirely lost and when, at last, on June 7 1946 the service was resumed, considerable plans for its future expansion had been prepared by the BBC. In 1949 a television transmitter with twice the power of Alexandra Palace was opened at Sutton Coldfield near Birmingham, to be followed soon afterwards by a similar transmitter at Holme Moss, near Huddersfield, and by 1952 the television service of the BBC had grown to the extent that three quarters of the population of the United Kingdom were within a satisfactory service area. With a total of seventeen television stations in operation and five more under construction, the national coverage provided by the BBC will, in the very near future, bring over 98 per cent of the population within its range—a density of coverage not exceeded anywhere else in the world.

But in the process of growing up the BBC has not been content to devote its technical resources solely to the completion of its coverage. It has made considerable strides in extending the sources of its television programmes, particularly in the field of outside broadcasting and much effort and ingenuity on the part of its research and engineering staff has been displayed to this end.

The first broadcast from outside this country took place in August 1950 when BBC engineers installed cameras and other equipment at Calais and set up microwave links across the Channel to convey the centenary celebrations of another remarkable engineering achievement—the laying of the first telegraph cable between England and France. From an engineering and a programme point of view it was a notable success, but the programme could not be transmitted over the French television network owing to the difference between the French and English systems.

Post-war growth of television in Western Europe had proceeded at a steady pace in spite of the conflicting views held on the different systems and the televising of the Coronation in 1953 to the Eurovision network represented an outstanding technical achievement.

The BBC has now, to all intents and purposes, achieved complete national coverage with a single programme, but to fulfil the terms of its Charter a second service is regarded as desirable and the BBC has pressed for the allocation of the whole of Band III for television purposes to enable it to proceed with this object in view. The allocation of channels for an alternative service is bound up to some extent with the more difficult problem of colour television and it is unlikely that any major decision can be made for another year or two. That a second service will be brought into being is certain. Colour television is inevitable, but whether or not the second service will, in fact, be colour is at this stage unpredictable.

Equally unpredictable is what lies beyond. What new wonders are in store for us when the BBC celebrates its television centenary cannot be imagined. The rate of progress during the past twenty-one years has been so remarkable that to hazard a guess as to the state of affairs no further ahead than the Silver Jubilee—a mere four years ahead—would seem unwise.

A Simple Shaft Digitizer and Store

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

The use of shaft position encoders to derive digital data from analogue instruments is discussed with particular reference to an extremely simple apparatus which has given very satisfactory service in daily use in a punched card data accumulating system. The encoder and the associated relay decoder-store are capable of being constructed in the laboratory or workshop.

THE advantages of logging and/or processing data in digital form are widely recognized and such methods have application whenever a multiplicity of observations is necessary to evaluate or specify a single operating condition. Applications of particular value are process data logging, process control and wind tunnel testing or engine bed trials. In many cases the rates of change of the test variables are sufficiently leisurely to permit quite simple electrical or electro-mechanical devices to be employed in the circuits, e.g. relays, solenoids and selector switches; having operating times of the order 2 to 20msec. High speed or 'one-shot' trials such as ballistic tests. rocket stand experiments or intermittent wind tunnel runs require a different approach and systems using components operating in the microsecond region are mandatory. Thus electronic systems with magnetic drum or tape storage or alternatively photographic and optical systems have been developed to meet these requirements. The technical virtuosity of these high speed installations has attracted well merited publicity, but it is the purpose of this article to show that quite elementary installations can provide a performance well suited to the requirements of a variety of industrial applications without recourse to the elaborate and expensive methods often associated with the phrase 'digital data processing'.

Broadly the high speed techniques tend to be less precise, e.g. many magnetic tape systems work to 7 bit accuracy, as against accuracies of 10 bits and upwards for slow speed systems; bandwidth considerations are clearly a limitation. The choice of approach and the optimization of a system for a given task can, and too often does, give rise to endless discussion of the niceties and elegancies of the many methods available. As a result wildly disparate estimates of the probable cost, complexity and reliability are current, to the detriment of the progress and prospects of a well established body of techniques already available. A selected bibliography at the end of this article shows how large a collection of material has accumulated in the past decade; too little of the material contains application data. The present article is aimed at filling that gap in respect of a particularly simple technique for obtaining data at modest speeds. It is hoped to show how simple the basic process of taking digital data can be made. More important perhaps it indicates that at the present time with an instrument industry bedevilled by the 'development contract' and large scale orders, it is practical and economical for new workers to produce useful data processing equipment for themselves with average laboratory or workshop facilities.

Servo Operated Instruments

Precision analogue methods of observing pressure, temperature, strain, position and the like commonly employ self-balancing devices such as strip chart recorders, potentiometer indicators and automatic Wheatstone bridges or else utilize transmission by magslips, synchros, step repeater motors or more complex r.p.c. mechanisms. From all these an analogue output in the form of the rotational displacement of a shaft is available. This is no more than a change of representation of the analogue quantity being observed; it is still essentially analogue in form and continuous. There may, of course, be many other changes of representation prior to the input to the recording device.

Analogue instruments may be broadly divided into two categories in terms of accuracy:

(1) Visual indicators and recorders of the self nulling type. These are limited in precision by the noise and drift levels at the servo amplifier input. Commercial chart recorders, for instance, are normally constructed to have an accuracy of written record of between 0·5 and 0·3 per cent of full scale above a full scale deflexion sensitivity of 5mV. Below this range a constant discrimination of typically 12μV is exhibited. However, when the discrepancies due to pen carriage drive arrangements are deducted, selected linear slidewires employed and due attention given to input noise levels accuracies of 0·1 to 0·2 per cent are quite feasible in the positioning of the main slidewire shaft for full scale ranges above 2mV. Hence digital representations quantizing to about 1 part in 1 000 will suffice. Typical full scale balancing times are 2 to 5sec.

(2) Position transmission links with inherent angular errors or ambiguities such as synchros or repeater motors. Where torque transmission is required the link elements may perform many revolutions in traversing full scale and the precision may be increased indefinitely at the expense of speed, by increasing the number of revolutions in the span, since angular discrimination is constant. Similar considerations apply to modern synchro links in which the elements have high inherent accuracies but torque transmission is achieved by the use of a velocity servo. Typical applications yield accuracies of 0·01 per cent of full scale and digital quantizing to 1 part in 10 000 is required: full scale traverse time is typically 20 to 50sec. Thus the bulk of the analogue instruments suitable for digital applications may be satisfactorily digitized with a quantum rate of about 500 counts/second. Normally digital data is sampled rather than recorded continuously and the sampling rate of a single channel will lie below 10 per second.

Digitizing

Digitizing an analogue quantity makes the basic assumption that it may adequately be represented by a sufficient number of discreet intervals. The choice of the number of intervals is clearly bound up with the accuracy requirement but depends upon less obvious factors such as the rate of change of the analogue quantity and subsequent processing and computational techniques. It is important when planning an installation initially that the required precisions and discriminations be clearly stated and under-

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stood. Over statement of accuracy requirements can lead to inordinate expense and complication of digitizing apparatus. Moreover precision is only adjustable stepwise e.g. by adding another binary digit which is impracticable beyond a well defined limit. Thus it is vital that the accuracy specification reflect real needs and not what 'it would be nice to have'.

Practicable Precisions

The limiting precision that can reasonably be expected from a single turn or single speed digitizing device is about 1 part in 1 200 unless physical bulk is unobjectionable. This adequately covers present self-balancing instruments and certainly all the transducers of engineering quantities in normal use. Two speed or multi-turn devices can be made with precisions of 0·01 per cent or even 1 in 32 000 which covers the range of familiar r.p.c. devices and the very few long-scale constant dicrimination devices available e.g. self-balancing weigh beams.

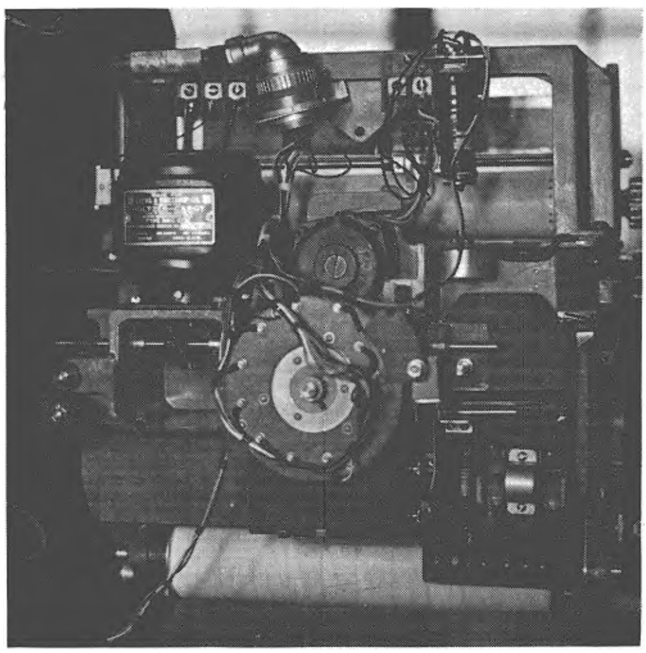

Fig. 1. Digitizer disk mounted on a commercial recorder

It is widely held that an accuracy once achieved immediately becomes the minimum requirement of the next specification sent to the instrument department. There seems however to be little prospect of bettering the above figures unless some other attribute, usually life or reliability, is to degenerate.

By far the bulk of digitizing devices is aimed at linear input/output relationships but curve fitting and function-generating types are practicable. Their precisions will generally be somewhat less than those quoted above especially if large slope changes exist in the functions.

Shaft Position Encoders

Shaft position encoders are one class of a wide variety of available or feasible methods of analogue-to-digital conversion[1 4]. Burke[5] lists 17 fundamental methods of conversion and the appended bibliography gives a fair survey of the techniques available. It appears to be true of analogue-to-digital convertors, as of transducers, that anyone can think of a new one—the difficulties of realization are often considerable. The shaft position digitizer is of especial importance, however, since it allows accepted and readily available instruments to be employed as sources of digital data.

Most recorders or transmission links are powered by a small servo motor which may perform between 5 and 500 revolutions in traversing full scale travel. It is seldom feasible to add much torque loading to the motors without loss of accuracy even if the inevitable increase in inertia loading can be accommodated in the stability margin of the loop. Consequently the shaft position encoder is usually driven at some point considerably geared down from the motor at which level the added friction torque can be accommodated. Where a variety of instruments are to be integrated into a unified scheme it is often more orderly to reduce all shaft motions to a single-turn-full-scale basis except where a genuine requirement for accuracies better than 1/1 000 of full scale exists, e.g. flow measurements. Fig. 1 shows a typical single turn digitizer mounted directly on the slidewire shaft of a potentiometer recorder.

Construction of a Digitizer

The digitizer used is basically a form developed at R.A.E. Farnborough[6] although commercial digitizers of

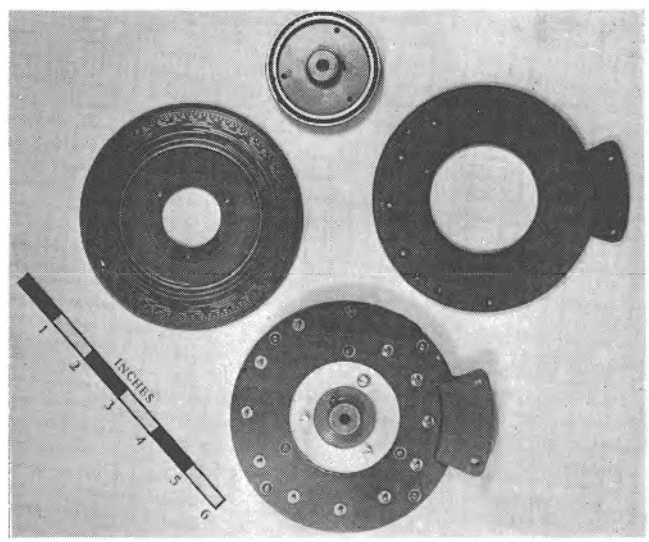

Fig. 2. Components of a digitizer disk

exactly similar principle are well known. As Fig. 2 shows, it consists essentially of a coded disk and a brush set. The coded disk is made by photo-engraving on copper, from a glass master negative, a pattern of copper lands arranged in annular tracks providing conduction with the brushes. The lands are separated from one another by filling the etched-out interspaces with an insulating resin (Araldite 103). The production of the glass master negative is discussed below under coding, but it may be noted here that masters exist already for most of the usual code forms.

The filled copper disk is cemented to a backing plate of good grade Tufnol and the commutating surface lapped by hand since lack of flatness in the copper engraving after cutting of bores, etc., makes local working essential. In practice machine grinding is continued down to +0·005in.

The brushes comprise 1/16in balls housed in silver steel bushes and held in contact with the commutating surface of the copper disk by means of small helical springs which also provide the electrical contact. The spring pressures are kept within 50g as measured on a relay springset gauge since this pressure has a marked effect upon the wearing properties of the device and also, together with other factors, upon the accuracy. The brush arrangement is seen in Fig. 3. The brushes are arranged in spiral fashion around the disk so as to avoid the difficulties of radial row grouping: the starting points of successive annular tracks being correspondingly staggered on the engraving.

binary code, because it contains such transitions as 511 (111 111 111) to 512 (1 000 000 000) is particularly prone to errors of reading when brushes are on the edge of such a transition (Fig. 4(a)). By rearranging the elements of the code so that only one digit changes at any transition such difficulties can be avoided. The rearrangement into this cyclic progressive form is not confined to the binary system

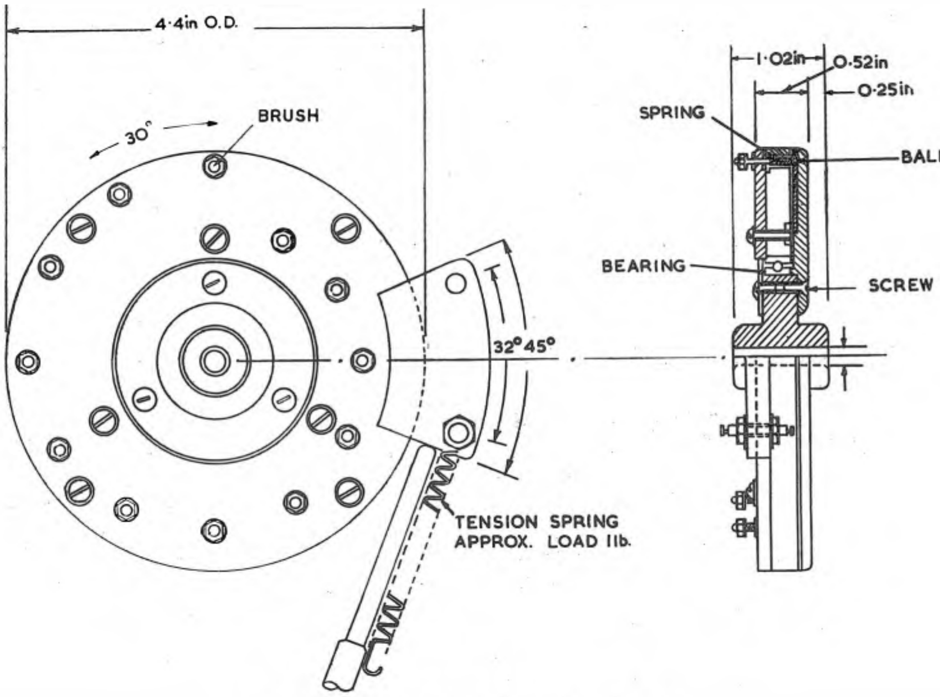

Fig. 3. General arrangement of digitizer disk

Apart from the engraving work, which was done commercially, the digitizers were made in the machine shop. Turning and boring tolerances were laid down to yield a concentricity of better than 0·005in since a cyclic error of 1 part per thousand arises if this is exceeded. The brush holders were positioned to $\pm 0{\cdot}0005$in radially and 6min of arc in relative angular position. The completed digitizer has an operating torque of about 2oz.in principally due to the radially-rigid uncaged ball-bearing used to support the brush carrier. The estimated cost of manufacture was about £20 per unit spread over a batch of 24 (excluding overheads).

Choice of Coding

Many factors affect the choice of a coding for digital readouts, but the use to be made of the data after reading has most weight. In many applications the data is destined for a digital computer and is almost always required to be stored in permanent form. Where monitoring or supervisory facilities are not the primary objective the use of decimal coding or its variants, e.g. binary coded decimal, has little attraction. In terms of simplicity of circuits, flexibility and ease of maintenance code forms based on the radix 2 are attractive. No real difficulty has been experienced in training technicians and engineers to read and think in binary. The advantages of simplicity and consequently enhanced reliability offset the educative effort involved.

Cyclic Progressive Codes

All engraved disk or commutator type digitizers must perforce employ a form of unambiguous coding or else special bush gear arrangements[7] to avoid ambiguities. The but can be carried out by simple logical operations for systematic codes of any radix[8]. A section of the cyclic progressive binary scale before staggering, is shown in Fig. 4(b). This code is also referred to as reflected binary or Gray code. Unfortunately it possesses no simple arithmetic so that additions, multiplications and digital-to-analogue conversions become unduly elaborate. Hence the cyclic progressive form is usually retained in a system only until

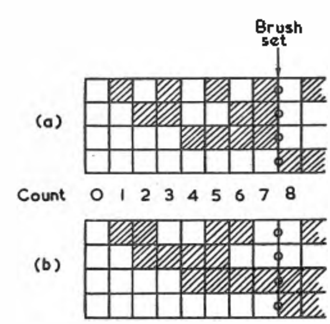

Fig. 4. Method of rearranging binary scale to avoid ambiguities

that stage beyond which ambiguities will not arise, at which point decoding into the more familiar form is carried out. Cyclic progressive codes, being merely a rearrangement of the code from which they are derived possess no additional error detecting properties, although by the addition of redundant bits an error detecting form may be derived[9].

Code Construction

A cyclic progressive form may be derived from a systematic code by simple logical operations on the 'pure' code form. If the digits of an integer in a pure code of radix R are denoted by $P_1, P_2, P_3 \ldots P_n$ in increasing order of significance then a cyclic progressive form may be derived in either of two ways

(a) *Parrallel Method*

The relation is simply $C_k = P_k$ if P_{k+1} is even or if P_{k+1} is odd $C_k = (R - 1) - P_k$ where $C_1 C_2 C_3 \ldots$ etc. are the cyclic digits in ascending significance. Table 1 shows the procedure for decimal ($R = 10$) and binary ($R = 2$). It will be seen that in both cases a multiple digit transition is reduced to a single digit transition.

TABLE 1

DECIMAL $C_k=P_k$, P_{k+1} even $C_k=9-P_k$, P_{k+1} odd		BINARY $C_k=P_k$, P_{k+1} even $C_k=2-P_k$, P_{k+1} odd	
P_4 P_3 P_2 P_1	C_4 C_3 C_2 C_1	P_4 P_3 P_2 P_1	C_4 C_3 C_2 C_1
0 3 7 8	0 3 2 1	0 1 1 0	0 1 0 1
0 3 7 9	0 3 2 0	0 1 1 1	0 1 0 0
0 3 8 0	0 3 1 0	1 0 0 0	1 1 0 0
0 3 8 1	0 3 1 1	1 0 0 1	1 1 0 1

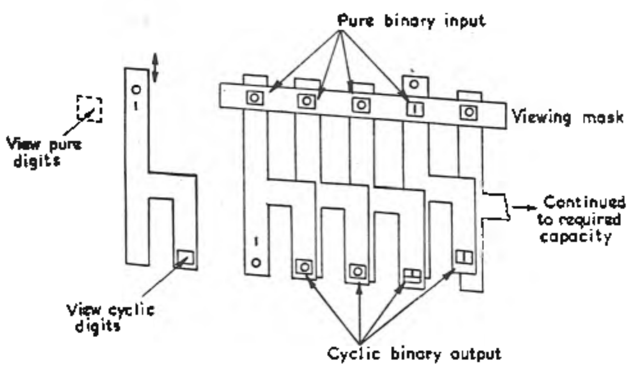

Fig. 5. Simple pure binary to cyclic progressive binary translating aid

(b) *Serial Method*

This is less apt for code construction but a similar principle is useful for reverting to the pure form. In this case the sum of the more significant digits preceding the cyclic digit being formed is examined and used as a discriminant.

i.e. $C_k = P_k$ if $\sum_{n-k-1}^{n} C_k$ is even and $C_k = (R - 1) - P_k$ if the sum is odd. Table 2 shows the procedure for binary numbers. The method is essentially serial since C_k cannot be determined until C_{k+1} is known.

The parallel method is to be preferred for code formation as all the requisite data is available in familiar easily checked form and errors invalidate only one element of the result which is likely to be more readily detected. The labour of constructing a binary code can be reduced by the use of a simple analogue mechanism shown in Fig. 5.

TABLE 2

	6	5	4	3	2	1
P_k	0	1	0	1	1	0
C_k	0	1	1	1	0	1
Parity of $\sum C_k$	0	1	0	1	1	-

PREPARATION OF MASTER NEGATIVES

Where masters are not already available two methods of procedure have been successfully applied:

(a) A large scale drawing, including a precise gauge length for guidance in subsequent reduction processes, is prepared: the accuracies required in construction, e.g. ± 3 minutes of arc necessitate very large scale drawings, and suitably large reduction equipment free from aberrations throughout the field are not common.

(b) A method well suited to engineering establishments was originated at R.A.E. Farnborough[10]. In this method the conduction/non-conduction patterns along successive radial stations are punched as hole/no-hole patterns in a succession of business-machine punched cards. The negative is then prepared by exposing a photographic plate mounted on a dividing head to the collimated slit image

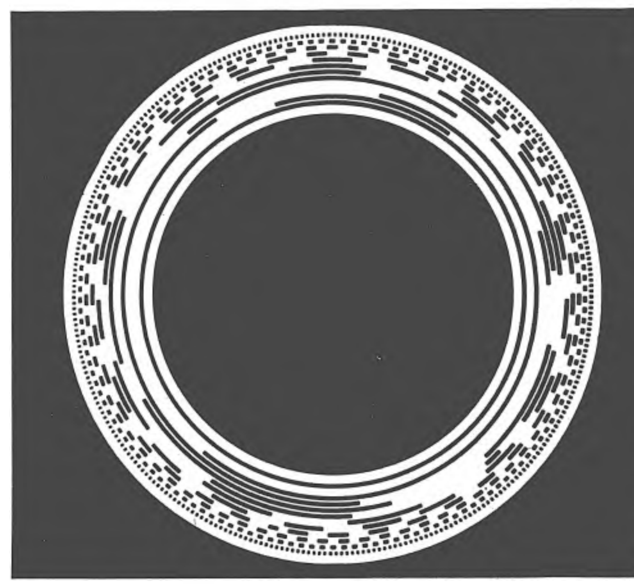

Fig. 6. Patterning of a 1 200 step cyclic progressive binary scale

of these cards, one at a time, the plate being advanced stepwise between exposures until a complete annulus has been formed. Subsequent processes are carried out using contact prints or reductions of this master. Fig. 6 shows the patterning of a 1 200 step cyclic progressive binary scale, the successive tracks of which have been staggered relative to one another. It may be noted that the five outermost tracks (less significant) are unchanged in this process since they repeat an integral number of times in the periphery of the scale. One benefit of the cyclic code form is that the repetition interval of the least significant digit is four integers as against two for pure binary so that the engraving of the least significant track is correspondingly eased. The preparation of the card pack for the master was originally done by hand but is a very suitable exercise for a digital computer.

Accuracy and Life Testing

Completed digitizers are subjected to a careful scrutiny after assembly: a primary test is to ensure that all the possible code combinations occur in the correct order with no drop-outs or skips to wrong combinations. This is simply done by visual inspection of a set of test lamps. It is generally found that a considerable scatter of the arc distances over which a given contact pattern is maintained occurs. Table 3 shows a typical extract from such a test record for a 1 200 count digitizer prepared by the punched card method.

TABLE 3

COUNT	ANGLE (deg.)	(min.)	ARC (min.)
171	51	15¼	22¾
172	51	33½	18¼
173	51	43¾	10¼
174	52	09¼	15½
175	52	28¼	19
176	52	44½	16¼
177	52	56¼	11¾
178	53	15	18¾
179	53	39½	24½
180	53	50¼	10¾

Mean Arc 17′

Since the device is quantized in its output this scatter does not alter the discrimination of the device which is limited to ± 1 count by the ambiguities of the brush pick-off provided that no cumulative errors occur such that N successive intervals sum to more than $N + 1$ nominal intervals (18′ in this instance is the nominal interval) or to less than $N - 1$ nominal intervals. Tests show that these limits are only just adhered to in the case of card constructed disks but rejections on this score have been rare even after wear. Most rejections have been due to excessive lapping of the filled copper disks resulting in the exposure of irregularly shaped lands arising from the under-cutting of the sharply defined outlines of the photographic pattern that occurs during engraving.

Wear Testing

The primary effect of wear in this form of digitizer is to form a groove along the path of the ball contacts. Such grooving will clearly extend the arc of contact of a ball with a flat land and this accounts for some of the scatter noted above: the spring pressure is obviously of prime importance here.

Wear apart, an empirical expression due to Bane (unpublished) relates the apparent increase in the electrical arc of contact to the mechanical length to the current and voltage at the contact as well as the spring pressure. The relation is of the form

$$\Delta L = a + bP + cV - dI$$

where a, b, c and d are positive constants; the expression appears to be valid for the values of voltage and current involved in relay readout of digitizer disks. The device described herein has sufficient diameter to make these additions to path length relatively unimportant, but they clearly form a limitation to the design of compact devices constructed on the same principle.

A preliminary batch of digitizers was subjected to oscillatory rotation of $\pm$ 150° and period about 4sec. Correct pattern succession was tested for at intervals of 50 000 reversals. The earliest failure was at 600 000 reversals and successful operation beyond 2×10^6 reversals was obtained before abandoning the test. Physical examinations showed that the majority of the wear occurred in the first few tens of thousands of reversals: simultaneously a soot-like powder was produced (christened 'Araldite oxide' but of uncertain parentage) chiefly deleterious in clogging the bores of the brush holders and causing loss of brush contact. The initial accumulation having been cleared no further deposits were noted and little apparent increase in ball track grooving occurred.

End of life as determined by coding errors outside $\pm$ 1 count arising, was in many cases due to tracking through of the filling compound due to excessive current commutation during checking. It is important in use that these devices are not allowed to commutate current: continuous commutation tests showed that life exceeding 50 000 reversals could not be expected under these conditions (relay currents of 20mA were involved), since the definition of pattern edges become blurred by arcing. It is believed that end of life in installed digitizers is usually brought about by arcing at the single ambiguities which may occur even with the disk at rest for readout.

Lives, expressed as the total number of quanta traversed, but not necessarily read out, significantly better than the above figures do not appear to have been achieved with either single or multiturn digitizers commercially available except where the reading brushes are arranged to be lifted out of contact with the coded plate(s) during traversing, with considerable increase in cost and complexity. It has been found economical and practical with the pattern described herein to regard the copper plate and its backing as expendable and to carry a stock of replacements—these cost about £5 each once a master negative is available. In an installation of about twenty such encoders two replacements have been made in 18 months of regular operation.

Decoding Operations

A fundamental decision in planning a data collecting system is the choice of the point at which multiplexing is to be carried out. The multiplexer or scanner may be designed to handle digital data in serial fashion or in parallel and the decoding operation may take place ahead of or subsequent to multiplexing. In the latter case it will be necessary to hold stationary the source of the data throughout the scanning period and this may involve locking up the observational instruments for a considerable period. In many instances such a 'flying blind' period may be dangerous or wasteful and efforts to minimize it are generally worthwhile.

Parallel data handling is no more expensive than serial methods when electro-mechanical speeds of operation are acceptable. The digitizer disk offers an essentially parallel output in the form of closures to a common line. Whether parallel or serial multiplexing is used it is useful for many reasons to complete decoding prior to the multiplexer. To minimize the lock-out period of the instruments storing may be conveniently added at the decoding stage so permitting scanning to be performed later, at a rate suited to the recording medium. The digitizer and the decoder/store then constitute a basic unit or channel that can be multiplied a suitable number of times to cover an installation independently of the multiplexing operation that follows. In this form it may be it could even be applied to existing installations with a minimum of interference to normal operation.

Decoder Logics

Again two methods of approach are possible: with the terminology of the earlier section one may represent them in tabular fashion as shown in Table 4.

TABLE 4

SERIAL MODE					PARALLEL MODE				
C_k	$+$	P_{k+1}	$=$	P_k	C_k	$+$	$\sum_{n-k-l}^{n} C_k$	$=$	P_k
0	+	0	=	0	0	+	0	=	0
1	+	0	=	1	1	+	0	=	1
0	+	1	=	1	0	+	1	=	1
1	+	1	=	0	1	+	1	=	0

It may be noted that in this case the summed digit parity method is the quicker as all the information needed is parallelly available. The serial mode requires P_{k+1} to be determined before P_k can be established. Fig. 7(a) shows a simple relay contact algebra to perform the operations described in tabular form: it will be seen that the required functions are identical for the two methods. Since the determination of the parity of the sum of the cyclic digits involves a serial chain of contacts throughout the length of the decoder the method is limited in practice by the current carrying capacity of such contacts.

Fig. 7(b) shows the serial mode of operation for a small number of digits; the pattern may be extended indefinitely. The initial connexion at the most significant end assumes a redundant 0 in the next, more significant place. This is clearly unobjectionable and must be assumed to initiate the decoding process. The parallel method is most suitable for reading out 'on-the-fly' or where speed is essential.

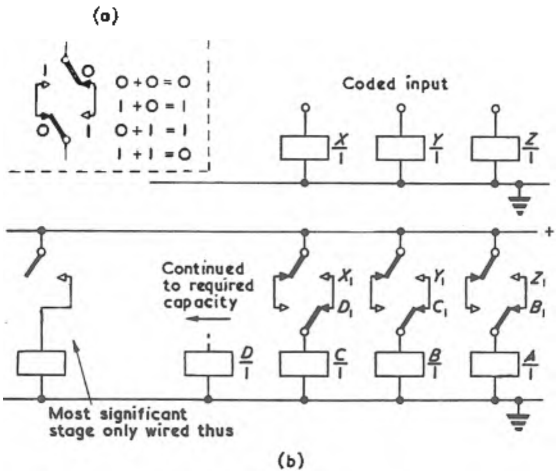

Fig. 7. Derivation of basic decoder circuit

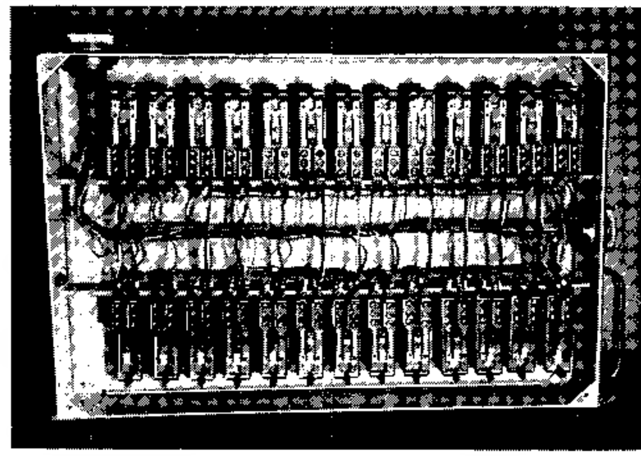

Fig. 9. Decoder/store for 10 bit cyclic progressive binary

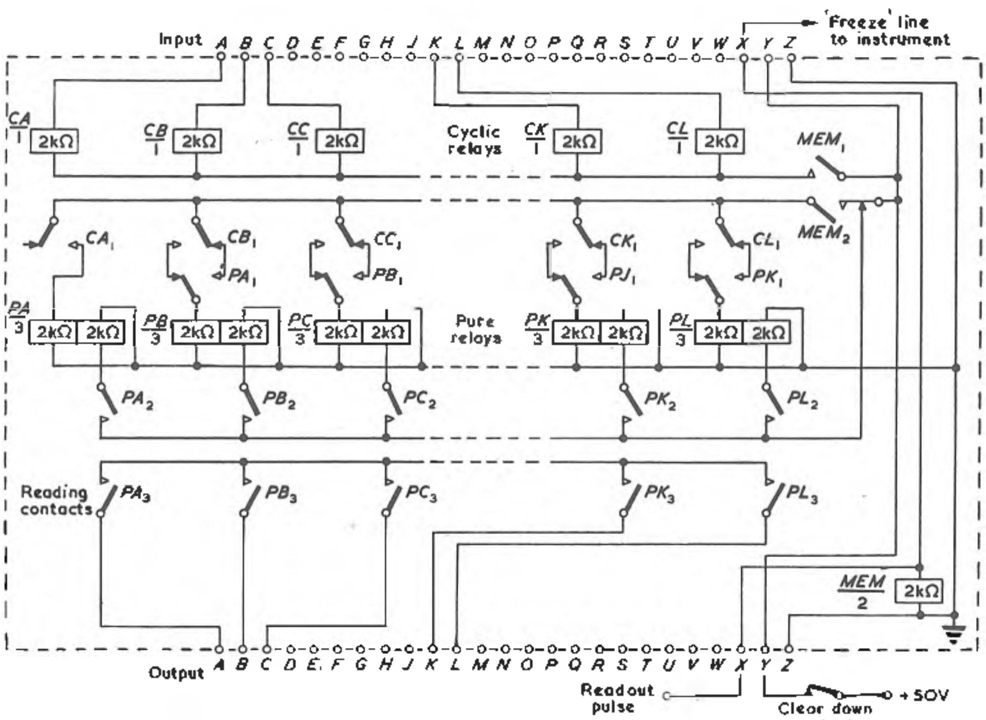

Fig. 8. General circuit of relay decoder

Decoder Circuits

Fig. 8 shows in outline the circuits of a practical decoder. In this the so called memory relay has been added. This relay is pulsed for a period sufficient to allow the serial decoding operation to take place completely. While it is operated the 'cyclic' relays are able to determine the state of the digitizer continuities and during this period the data source must be held stationary or 'frozen'. This is achieved by operating a further relay, in parallel with the memory coil, whose contacts are arranged to halt the servo-motors, e.g. by open-circuiting the reference winding of a two-phase type. The frozen period is the relay operating time multi-

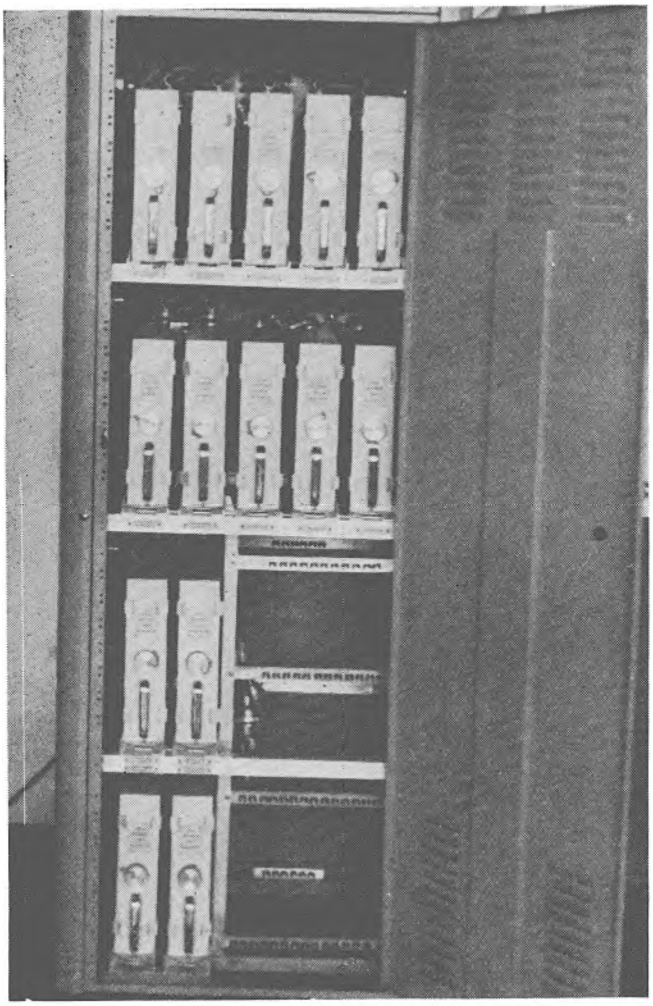

Fig. 10. Typical installation of decoder/stores

plied by the number of bits, i.e., about 200msec for a 10-bit instrument.

As the memory relay falls back the decoded information is held by the action of the self hold contacts on the 'pure' relays which lock up on their second coils. The pure binary relays then retain the digital data as long as the supply is maintained. The cyclic relays are cleared down and do not follow subsequent movements of the digitizer: the storage period is terminated by a brief interruption of the power supply ('clear-down' contact).

Decoder Construction

In order to keep costs low and maintainability high the decoder/store was constructed from standard K3000 relays (P.O. major). The open construction shown in Fig. 9 has given extremely reliable service even during accelerated life testing when deliberate accumulations of dust were permitted prior to and during testing. Ten years equivalent normal operation was applied without failures: in-service failures of a batch of 18 over 1½ years have totalled three, only one of which was due to contact defects. No errors in recorded data have been attributed to decoders during scheduled running involving over one million decoder operations. The decoders were produced in the laboratory at a cost of about £50 each (time and materials).

The dimensions of the decoder/store are somewhat large (12in × 18in × 3in) and could no doubt be achieved in a much smaller volume. Such compression has no intrinsic merit and will certainly reduce serviceability.

Data handling equipment should essentially be accessible and Fig. 10 gives an impression of the entire decoding/storage apparatus for a 14 channel readout being installed in a supersonic wind tunnel control area. It is housed in a 7ft, 22½in relay rack: the space in the lower right-hand region will house the particular form of multiplexer required to sequence results from the stores into pre-arranged locations on punched cards. The rack and the associated card punch form the complete data taking installation.

Conclusions

A digitizer-decoder-store block suitable for building into digital data handling installations has been described and the construction processes outlined. Such blocks are capable of being produced in the laboratory or workshop for under £100 each. They are reliable and accurate in service and cost little to maintain.

Acknowledgment

The author is indebted to the Chief Executive of the Aircraft Research Association for permission to publish this article.

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VIBRATION MEASUREMENTS

Employing Ba Ti O_3 Accelerometers

By Jens T. Broch*, Dipl.ing.E.T.H.

The possibility of measuring the properties of mechanical structures by means of electrical methods is generally appreciated. However, the introduction of piezo-electric ceramics such as barium titanate opens new possibilities in miniaturizing the mechanical-electrical transducers, thereby reducing the 'back effect' of the transducer upon the test-object, and making measurements possible in places where space is limited.

Furthermore, employing 'automatic measuring technique' a complete frequency-amplitude diagram (spectrogram) of complex vibrations can be obtained in a few seconds. A consideration of the basic ideas behind such systems as well as a description of a commercially available system is given.

IN modern design and maintenance of mechanical structures it is of great importance to be able to determine the magnitude and frequency of vibrations occurring at different places in the structure when it is subjected to dynamic forces. Not only is it possible from the measurement of vibrations to apply an effective insulation technique and thus reduce the external effect of the vibrations to a minimum, but quite often the measurement of vibrations enables the detection of faults in the motivating unit to be made.

Recent development of mechanical-electrical transducers, together with reliable and sensitive electronic measuring equipment has made it possible to measure the dynamic properties of mechanical structures with a high degree of accuracy. Such measuring techniques have become an important factor in development.

In the following a measuring system based on the use of piezo-electric accelerometers will be described, and some examples of their application to actual industrial problems given.

There are three variables in a vibrating system which are of interest: the acceleration, the velocity and the displacement of the system.

These variables are connected through the well known relationships:

$$v = (\partial s/\partial t) \text{ and } a = (\partial v/\partial t) = (\partial^2 s/\partial t^2)$$

where s is the displacement, v the velocity and a the acceleration.

Using electronic equipment in connexion with a vibration pick-up it is not normally essential which of the three quantities are actually measured by the transducer, because the remaining quantities can be found by electronic differentiation or integration.

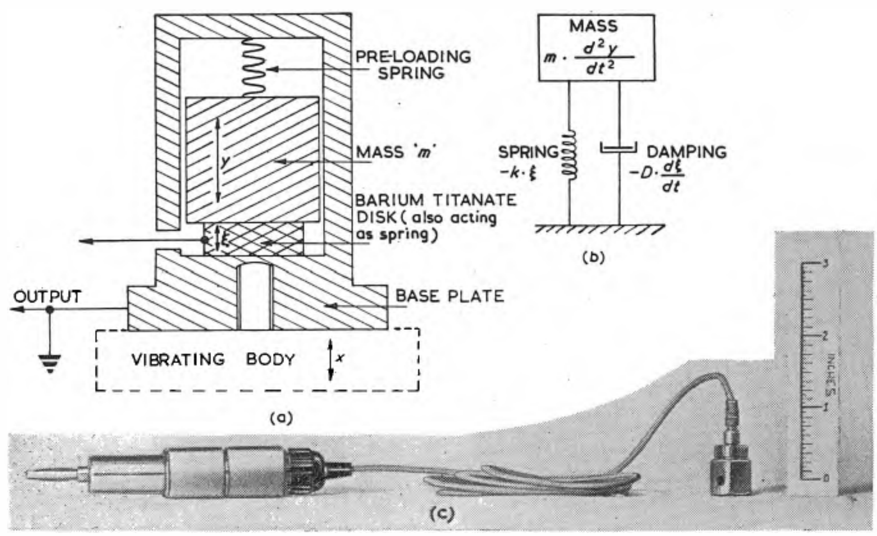


Fig. 1. (a) Principle of construction of modern wide range accelerometers
(b) Equivalent mechanical diagram of an accelerometer
(c) A commercially available accelerometer (Brüel & Kjær, Type 4329)

However, due to the fact that the output from an accelerometer is proportional to the force applied ($F = m \cdot a$) this type of pick-up will be the most convenient one to use in many vibration-measurements.

Fig. 1 shows the principle of operation of a piezo-electric accelerometer.

It consists basically of a mass-spring-system, the movement of the mass m producing a strain e in the piezo-electric material. The strain e in turn produces a potential difference in the material, depending upon the material's piezo-electric properties.

To give a clear understanding of the operation of the accelerometer the equation of motion of the mass m should be considered.

If the absolute movement of m is denoted by y, and the movement of the 'base' is denoted by x the relative movement of the mass ξ is:

$$\xi = y - x$$

* *A/S Brüel & Kjær, Denmark.*

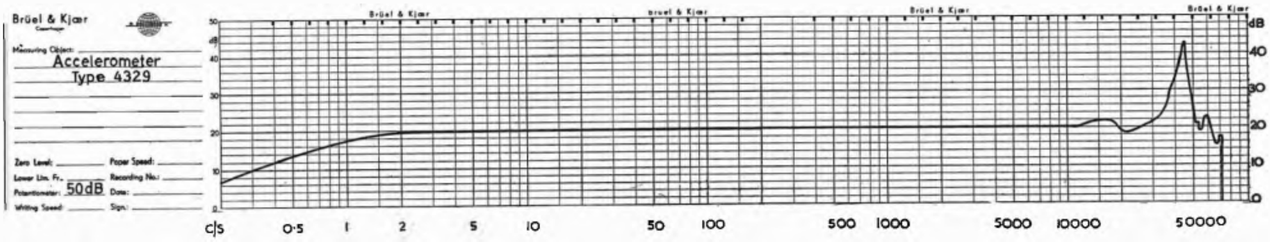


Fig. 2. Typical frequency characteristic of a wide-range accelerometer (Brüel & Kjær, Type 4329)

The force acting on the spring will thus be:

$$F_s = -k \cdot \xi$$

where k is the 'spring-constant'.

The damping force F_d (see Fig. 1(b)) is equal to

$$F_d = -D \cdot (d\xi/dt)$$

and the total external force moving the mass m must be:

$$F_{ex} = m \cdot (d^2y/dt^2)$$

$$m \cdot (d^2y/dt^2) = m(d^2\xi/dt^2 + d^2x/dt^2) = -D\,(d\xi/dt) - k \cdot \xi$$

or

$$-(d^2x/dt^2) = (d^2\xi/dt^2) + (D/m) \cdot (d\xi/dt) + (k/m) \cdot \xi$$

By assuming the system to be subjected to a sinusoidal vibration $x = x_a \cdot \cos \omega t$ a solution to this differential equation is:

$$\xi = x_a \frac{\omega^2}{\sqrt{[(k/m - \omega^2)^2 + (\omega(D/m))^2]}} \cos(\omega t - \phi)$$

The natural frequency of the mass-spring-system is determined by:

$$\omega_0 = \sqrt{(k/m)}$$

and for frequencies much lower than this value:

$$\xi = (m/k) \cdot \omega^2 \cdot x_a \cdot \cos(\omega t - \phi)$$

assuming the damping term to be negligible at these lower frequencies.

Comparing this expression with

$$(d^2x/dt^2) = -x_a\omega^2 \cdot \cos \omega t$$

it is clearly seen that the strain e which is proportional to ξ will also be proportional to the acceleration of the vibration considered.

The output voltage from the pick-up will therefore be proportional to the acceleration of the vibrations to which it is subjected, as long as the accelerometer is operated below its mechanical resonance frequency. A typical frequency characteristic of an accelerometer employing barium titanite as piezo-electric material, Fig. 1(b), is given in Fig. 2.

The equivalent diagram of a piezo-electric accelerometer will be as shown in Fig. 3, with a generator, whose output voltage is directly proportional to acceleration, in series with a capacitor C_a, and loaded with the insulation resistance of the accelerometer itself, R_a.

The low-frequency cut-off of the accelerometer is determined by these constants. Loading the output of the accelerometer with the input impedance of an electronic amplifier reduces the value of R, and raises the low-frequency cut-off of the system. It is therefore of importance to use an amplifier with a high input impedance, as the first 'link' in the amplifying and measuring system.

This system may consist of a valve-voltmeter only, or of more complicated amplifiers in connexion with filter networks, for the analysis of the vibration frequency spectrum.

Several measuring systems have been developed which are more or less convenient for the different types of vibration measurements and analyses met in practice.

A 'standard' type of measuring arrangement offering the advantage of obtaining automatically a complete frequency-amplitude diagram of the vibrations considered, is shown in Fig. 4.

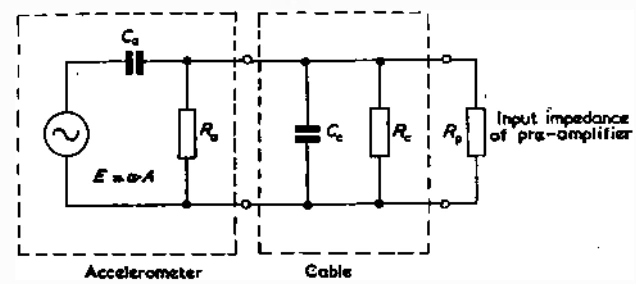


Fig. 3. Equivalent electrical diagram of an accelerometer when connected to the measuring instruments

Basically the system consists of an accelerometer, a pre-amplifier which should load the pick-up as little as possible, a selective amplifier and a direct-writing level recorder.

The pre-amplifier is placed close to the measuring point to avoid the use of a long connecting cable between the accelerometer and the pre-amplifier, as this in many cases influences the sensitivity of the set-up and invalidates the accelerometer sensitivity data given by the manufacturer.

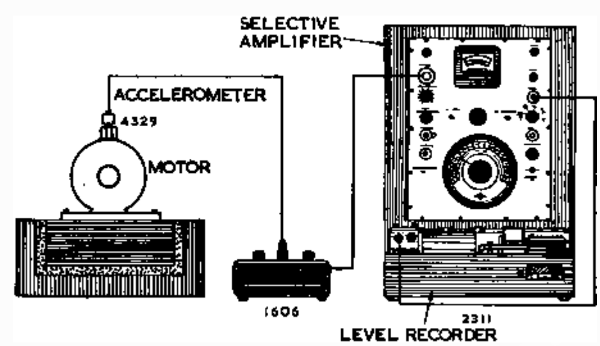


Fig. 4. Measuring arrangement for vibration measurements on a motor

From the equivalent diagram of Fig. 3 it is seen that a capacitive voltage division takes place between C_a and the cable capacitance C_c, the input voltage to the pre-amplifier thereby being reduced.

The accelerometer shown in Fig. 1(c) has an internal clamped capacitance of approximately 700pF. The relationship between the length of cable and the decrease in sensitivity is shown in Fig. 5 for a cable of approximately 60pF/m. Furthermore the low-frequency cut-off of the set-up will also be influenced, and using the symbols of Fig. 3, the resulting cut-off frequency will be:

$$f_0 = 1/(2\pi RC)$$

where C is the parallel combination of C_a and C_c, and R is the parallel combination of R_a, R_c and R_p.

This effect may be used to extend the measuring range

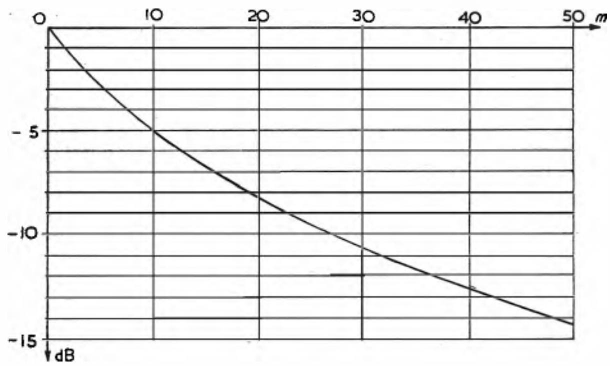


Fig. 5. Typical curve showing the reduction in sensitivity when an accelerometer is connected to the pre-amplifier via a connecting cable

downwards when it is desired to measure at extremely low frequencies and when a loss in sensitivity can be tolerated. Normally, however, the effect of the cable upon the frequency characteristic will be of no importance, and only the loss in sensitivity must be taken into account.

The main function of the pre-amplifier is thus to 'transform' the high output impedance of the accelerometer to a lower impedance so that any desired length of cable can be employed between this amplifier and the analyzing and recording arrangement.

Quite often, however, the pre-amplifier also contains the integrating networks necessary for the measurement of velocity and displacement of the vibrations considered. When only pure sinusoidal vibrations are being measured it is not strictly necessary to employ these networks because the velocity and the displacement can be found mathematically from the measurement of acceleration as:

Velocity,

$$v = \int_0^t a dt = A \int_0^t \cos(\omega t) = (A/\omega)\sin(\omega t - \phi_1)$$

and

Displacement,

$$= s \int_0^t\int_0^t a dt dt = \int_0^t v dt = (A/\omega)\int_0^t \sin(\omega t - \phi_1) dt$$

$$= -(A/\omega^2) \cdot \cos(\omega t - \phi_2)$$

On the other hand, when measurements are carried out on vibrations having continuous frequency spectra neither the velocity nor the displacement can be found mathematically purely from acceleration measurements, and it is therefore practical to use a pre-amplifier with built-in integrating networks.

The selective amplifier used in the set-up of Fig. 4 employs a series of band-pass filters, each filter having a bandwidth of ⅓ octave and a heavy attenuation of frequencies outside this pass-band (see Fig. 6).

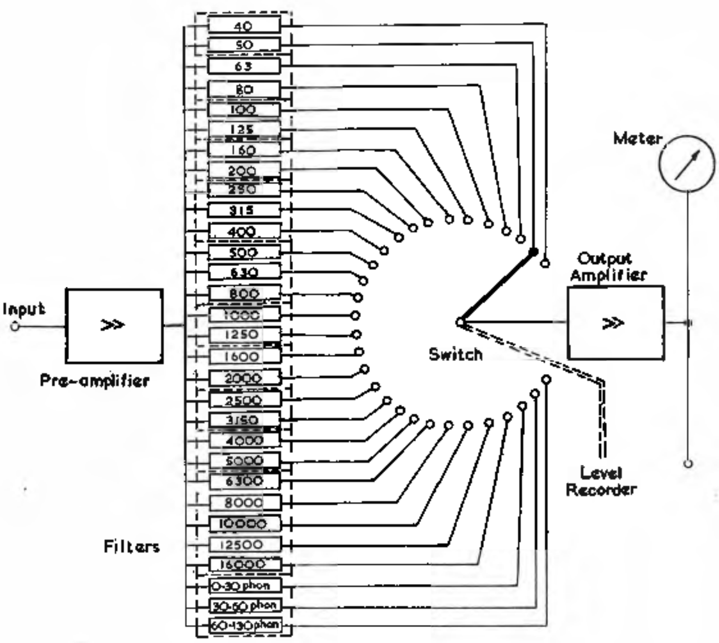
To reduce the effects of transients produced when the instrument is switched from one filter to the next the input circuits of the filters are paralleled and the switching arrangement included in the filter output circuit.

The filter switch can be mechanically coupled to the motor-drive in the level recorder, Fig. 4, whereby the switching can be carried out automatically. By selecting a proper paper speed on the level recorder the paper can be run in synchronism with the frequency sweep of the analyzer filter switch. It is then possible to use pre-printed, calibrated recording paper, whereby a written, calibrated frequency amplitude record of the vibrations being measured can be obtained, (see Fig. 8).

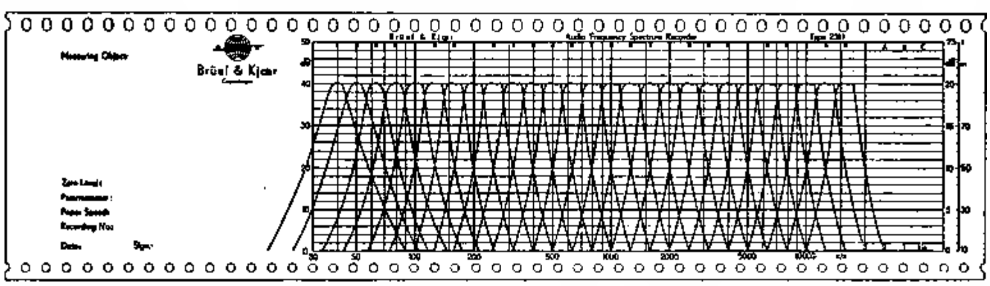
The operation of the level recorder used in the measuring arrangement described is based upon the servo-principle, (see Fig. 7).

When the magnitude of the voltage applied to the input terminals is changed a current will flow through the driving coil of the writing system, thus moving the stylus, which is mechanically coupled to the input potentiometer. By the movement of the stylus the voltage delivered from the potentiometer to the a.c. amplifier will be altered until a stable condition is regained. In this way it is possible to obtain different ranges of voltage variations which can be recorded, by employing different input potentiometers.

To illustrate the use of vibration measurements in modern industry a typical application of this measuring technique is given below. A measuring set-up as shown in Fig. 4 is used to detect mechanical faults in rotating machinery.



(a)



(b)

Fig. 6. (a) The selective amplifier used in the measuring arrangement of Fig. 4 (Brüel & Kjær, Type 2109) (b) Filter curves for the selective amplifier

This type of measurement can be used for production control of motors and generators, and the only requirement to be satisfied is that each test object must be mounted in exactly the same way. A typical example of spectrograms obtained from tests on small a.c. motors is shown in Fig. 8. In Fig. 8(a) the spectrogram of a 'normal' motor is shown, this may be compared with Fig. 8(b), in which the result grams greatly reduce the time and work spent on finding the correct insulation technique in each case.

Examples of spectrograms obtained from measurements of vibrations on the floor close to where a motor has been mounted are shown in Fig. 9. Curve A was measured with the motor set-up on solid supports and curve B when it was suspended on vibration absorbers.

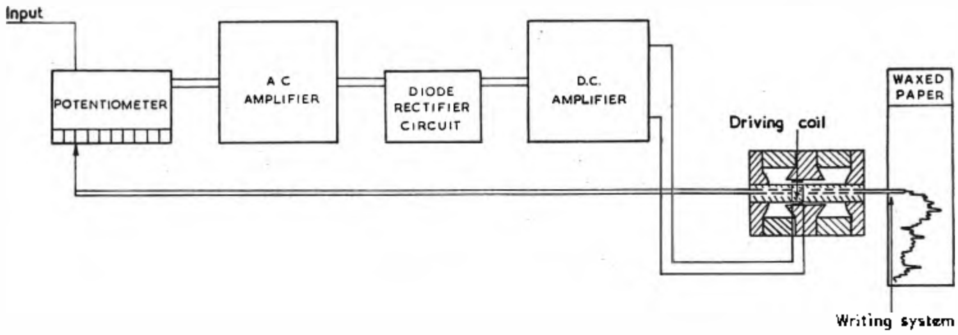


Fig. 7. The level recorder used in the measuring arrangement of Fig. 4 (Brüel & Kjær, Type 2304)

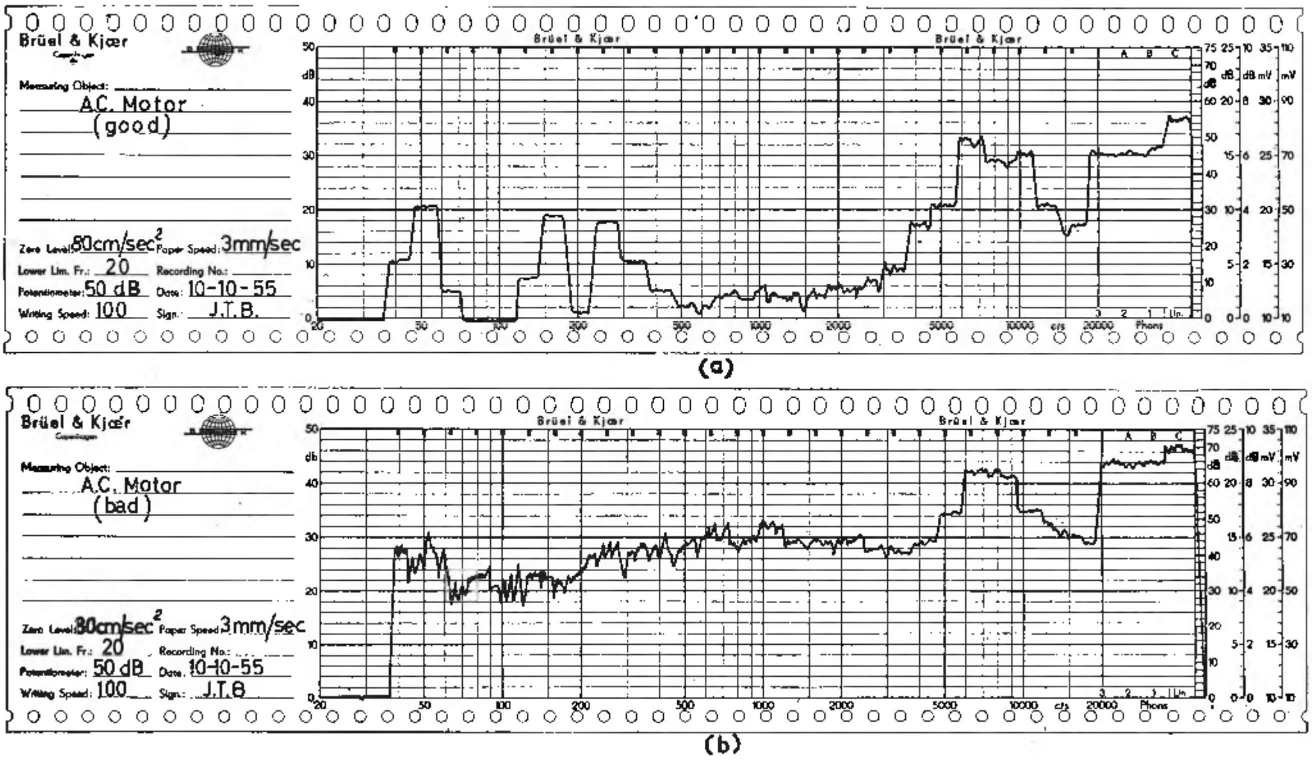


Fig. 8. Spectrograms obtained from vibration measurements on small a.c. motors

(a) Vibrations from a 'normal' motor

(b) Vibrations from a 'faulty' motor of the same type

obtained from a motor with a faulty bearing can be seen. Sound spectrograms were also taken of the noise radiating from the two motors, but no significant difference was found between these curves.

Another important type of vibration measurements is the measurement of effectivity of vibration insulation materials and installations. It is well known that vibration will irritate human operators, the irritation effects being dependent upon both the frequency and the magnitude of the vibrations. Furthermore vibrations may cause damage to expensive machinery or buildings, unless attention is given to insulating and damping their effects.

Insulation may sometimes be a rather difficult problem, and the use of automatic recording of vibration spectro-

The examples of vibration measuring technique given here are meant only as illustrative examples of the use of a measuring set-up as shown in Fig. 4, and there are, of course, many other fields of technical importance such as the vehicle, ship and aircraft building industry where vibration measurements are applied.

The evaluation of the measuring results shown in Figs. 8 and 9 is based on relative measurements, and the absolute calibration of the measuring arrangement is then of less importance than when accurate absolute measurements of vibration levels are to be taken. In cases where it is desired to know these levels the frequency response and accuracy of each instrument must be considered or the complete set-up must be calibrated as a whole.

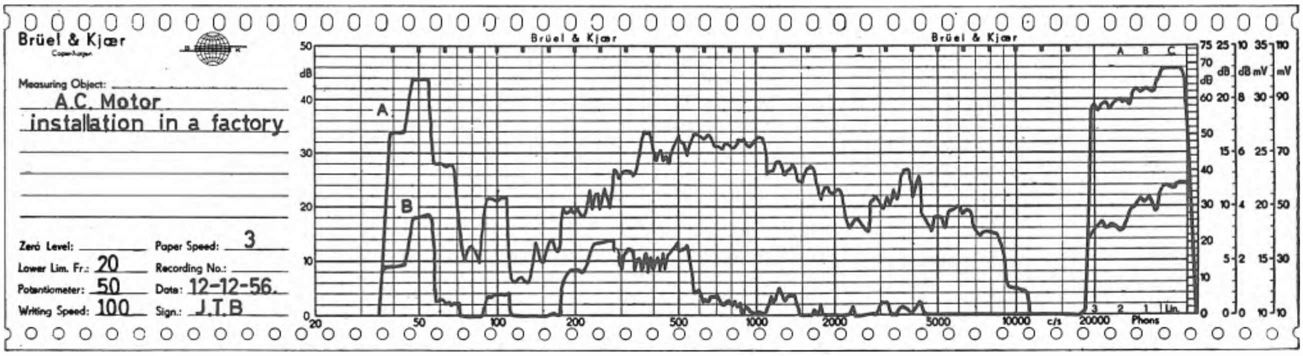


Fig. 9. Spectrograms showing the effect of vibration insulation

Curve A *was obtained without employing a vibration insulation technique and curve* B *was measured when vibration absorbers had been installed*

The pre-amplifier used in the set-up of Fig. 4 is equipped with a 'vibrating table', and an absolute calibration of the measuring arrangement can be made at an acceleration level of $1g = 981\text{cm.sec}^{-2}$. The frequency characteristic must either be calculated on the basis of the frequency characteristic given by the manufacturer for each separate instrument in the measuring system, or measured under the conditions actually present in practice. If the resulting frequency characteristic is calculated, however, the effects of connecting cables must then also be considered.

Acknowledgments

The author wishes to thank the management of A/S Brüel & Kjær for their kind permission to publish this article.

Components Symposium at Malvern

The first international symposium on electronic components was recently held at the Royal Radar Establishment at Malvern and its purpose was to review and discuss component development, new materials and new techniques for components operating under the severe environmental conditions imposed by their use in supersonic aircraft and guided missiles. Under the chairmanship of Brigadier J. D. Haigh (Director of Electronics Research and Development Telecommunications Ministry of Supply) some twenty-seven papers were presented, covering practically every type of component, with the exception of the valve and cathode-ray tube, for use in modern electronic equipment.

The full proceedings of the Symposium, together with the discussions of the papers will be published by the end of the year.

The Laboratories of the Technical Services Division were open to the delegates attending the symposium and on view were the latest examples of component development, constructional techniques and environmental testing methods..

Among the facilities for environmental testing is a stratospheric test house which is designed to simulate, in the laboratory, any known climatic condition to which components or equipment in an aircraft may be subjected.

The test house has a temperature range of from $-85°$C to $+150°$C with a pressure range of 20lb/in^2 absolute to 75 000ft altitude and humidity range of 30 per cent to saturation.

The particular features provided are:—

(a) Short times taken to obtain $-55°$C or high temperatures.
(b) Approximate simulation of the flight plan climatic environments of the equipment bay in supersonic aircraft.
(c) A wide range of combined temperature, humidity and altitude.

It is the only known chamber in the U.K. capable of providing these facilities.

Temperature is reduced by two coolers inside the vessel. Temperatures down to $-40°$C are obtained by circulating through one cooler trichlorethylene when is pumped from a storage vessel held below $-40°$C by a 75 h.p. ammonia plant. Temperatures below $-40°$C are obtained by direct expansion of arcton 3 from a 35 h.p. compressor in the second cooler.

Heating is obtained from a 66kW three-phase heating bank of sheathed elements. Humidity is obtained from steam produced by a 30kW electrode boiler; humidities below local ambient are obtained by first condensing out all the moisture in the vessel and then adding as required.

A constant mass flow air circulation is provided within the vessel by a 40in diameter two-stage axial flow fan driven by a 5 h.p. variable speed motor. A ducting and plenum system in the vessel ensure adequate uniformity of temperature in the test space. Flow rates for the fan range from $33\,000\text{ft}^3$/min at altitude to $3\,000\text{ft}^3$/min at sea level.

Pressures in the test space are obtained by a 35 h.p. 2-stage rotary vacuum pump remotely housed in a sound proofed room.

A further facility of the Laboratories at R.R.E. Malvern is an Automatic Component Testing Equipment (ACTE) for the testing of batches of resistors up to 1 000 in number over a wide range of climatic conditions.

It consists of a test chamber 5ft × 6ft × 5ft into the top of which fits a jig-block carrying the components to be tested. Temperatures between $-75°$C and $+100°$C can be provided and held to within $\pm 2°$C.

About 5°C the humidity can be controlled between 10 and 95 per cent to an accuracy of 2 per cent.

The equipment can test 1 000 resistors of values between 10Ω and $1\text{M}\Omega$, one value at a time. The resistors are divided into four groups of 250, and each group is capable of being differently a.c. loaded to a maximum of 250W.

Resistor measurements, which are made sequentially, each take 18sec. The switching of each resistor into the measurement circuit is controlled by the recording chart via a photo-electric cell. Measurements are made by means of a self-balancing Wheatstone bridge which automatically records the readings as a percentage deviation from standard on a special chart recorder, using a single stamp marker. Each series of readings is displaced by 1/16in from the previous one, allowing space for approximately 40 measurements of each resistor on the chart, which is 200ft in length.

Resistor measurement accuracy is within $\frac{1}{4}$ per cent.

The pan-climatic cabinet is heated by rod heaters, totalling 7kW which are brought into operation as required.

Cooling between 100°C and 38°C (approx.) is by means of mains water and between 38°C and 15°C by a $\frac{1}{2}$ h.p. refrigerator. Below 15°C a 5 h.p. refrigerator is brought into operation. The appropriate means of cooling is automatically determined by thermostats in the pan-climatic cabinet. Two atomizers spraying distilled water into the cabinet introduce the required humidity. Reduction in humidity is effected by condensing out the water vapour using the cooling system. An air circulation of 10ft^3/sec is maintained by two fans.

The automatic programming and control equipment obtains instructions from a pre-punched 5-hole Creed tape which moves 3in/h past the transmitter head. A number of pre-set control circuits are brought into operation. These include 15 pre-set climate panels, resistor loading power units and resistor measuring circuits. The latter are not effective until standard conditions (20°C, 75 per cent r.h.) are achieved in the cabinet.

The maximum length of test is limited only by temperature and voltage recording charts and is 120 days. Various interlock and inhibitor circuits are included in the equipment.

Climatic control is achieved by using thermistors as wet and dry bulb thermometer elements which operate the pan-climatic gear through an electronic bridge circuit and amplifier. Potentiometers in the pre-set climate panels form another bridge arm. The bridge is balanced by obtaining the "pre-set" climate in the cabinet.

A Wide Band Analogue Multiplier Using Crystal Diodes

and its Application to the Study of a Non-linear Differential Equation

By Michael E. Fisher*, Ph.D.

The detailed design, construction and alignment of an inexpensive wide band electronic analogue multiplier are described. Four-quadrant operation is obtained by the quarter-squares principle and each non-linear (squaring) unit employs twelve germanium diodes in the feedback loop of an operational amplifier (as in MacKay's designs). These units have an asymmetric quadratic characteristic and produce less than 1° phase shift at 50kc/s. The output errors of a multiplier are less than 0·3 per cent of the maximum output.

Two multipliers have been used with a high-speed repetitive analogue computer to investigate a non-linear differential equation which depended on a number of parameters.

THE need for a cheap, accurate and reliable electronic device to form a voltage proportional to the product of two other voltages has been felt for many years. Numerous devices to perform multiplication have been invented[1], but it cannot be said that the problem has yet been solved ideally. If electronic analogue computing is to progress to handle more difficult non-linear problems, reasonably accurate multipliers must be readily available.

Most multipliers that have been developed so far suffer from relatively severe frequency limitations which make them unsuitable for use in repetitive computers. Even at low repetition rates there must be ample bandwidth 'in hand' if the computing errors due to the frequency limitations (first evaluated by Macnee[2]) are to be avoided.

Some recently developed multipliers[3] based on the use of specially designed electron-tubes promise better frequency response, but they are relatively costly.

The present article describes the design, construction, alignment and application of a multiplier which, at moderate cost, provides a good frequency response and reasonable accuracy.

The principle of quarter-square multiplication[4] is used. The squaring is performed by networks of biased germanium diodes in the feedback paths of standard computing amplifiers. With this arrangement the phase shifts due to the diode capacitances may be compensated so that the non-linear units of the multiplier display negligible phase shift (less than 1°) up to 50kc/s. The design is essentially similar to a circuit described recently by D. M. MacKay[5], but it is thought that a more detailed discussion of the procedure of design and alignment may be of value, and the present multiplier has been designed to achieve greater accuracy.

By using 12 diodes in each squaring unit (i.e. 24 per multiplier) the maximum errors have been reduced to less than 0·3 per cent of the maximum output. (It is in the nature of the quarter-squares principle that the zero error may be of the same absolute magnitude as the errors elsewhere on the characteristic: since a product is normally integrated in practical computation, this is not too serious).

In addition to 24 germanium diodes each multiplier uses five computing amplifiers; in the present case this means ten valves. The power supplies must be stabilized but may be of orthodox design. All components are standard and readily available.

Design of the Multiplier

Fig. 1 shows the block diagram of the multiplier. Five amplifiers are used; numbers 1 and 4 function merely as sign-changers while 5 is an adder. Amplifiers 2 and 3 (and their associated feedback and biasing circuits) constitute two similar non-linear units for which the output voltage is a quadratic function of the input voltage:

$$V_0 = aV_1^2 + bV_1 + c \quad \ldots\ldots\ldots\ldots (1)$$

x and y are the input voltages to the multiplier; they are fed directly to the second non-linear circuit (amplifier 3) whose output is

$$V_{0,3} = a(x + y)^2 + b(x + y) + c \quad \ldots\ldots\ldots (2)$$

The first non-linear circuit (amplifier 2) is fed with x and $-y$ so its output is

$$V_{0,2} = a(x - y)^2 + b(x - y) + c \quad \ldots\ldots\ldots (3)$$

When equation (3) is subtracted from equation (2) the result is

$$4axy + 2by.$$

If the correct fraction of the input y is subtracted from this sum only the product term $4axy$ remains. These subtractions are carried out in amplifier 5 whose output is thus proportional to the product xy.

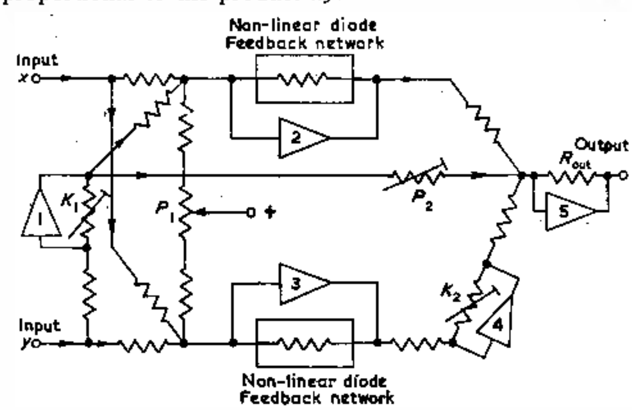


Fig. 1. Block diagram of the multiplier

All input and feedback resistors except P_2 and R_{out} may have the same value; this was chosen to be 100kΩ. The feedback resistors of amplifiers 1 and 4 were made adjustable over a 10 per cent range so as to permit accurate alignment of the multiplier (see below). The appropriate value of P_2 depends on the parameters a and b of the non-linear circuit. In the present case P_2 was variable between 40 and 60kΩ.

The value of R_{out}, the final feedback resistor is a matter of convenience (limited only by the linear range of amplifier 5) since it merely determines the scale factor of the output. If the time integral of the product xy is required, R_{out} may be replaced by a capacitance which converts this stage into an integrator. Small air trimmers (30pF maximum) are connected across each input and feedback resistor in order to eliminate phase shifts (see below).

Amplifiers

The choice of amplifier is determined by the intended use of the multiplier and does not otherwise affect the design of the multiplier. In the present case the multipliers were constructed for high-speed repetitive analogue com-

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putation. Accordingly the simple two-stage computing amplifier shown in Fig. 2 was used[5]. This is a high frequency compensated d.c. amplifier with a gain of about 70 and a half-value bandwidth of about 750kc/s. An adding circuit including this amplifier has a bandwidth of more than 7Mc/s (after adjustment of trimmer capacitances) and an output impedance of less than 20Ω. The output may range between plus and minus 15V.

Five stabilized power lines supply the multiplier amplifiers (and the rest of the computer). The d.c. level of an amplifier output may be set by adjusting the cathode bias potentiometer. No drift compensation[6] was included but for applications at lower speeds this would be very desirable.

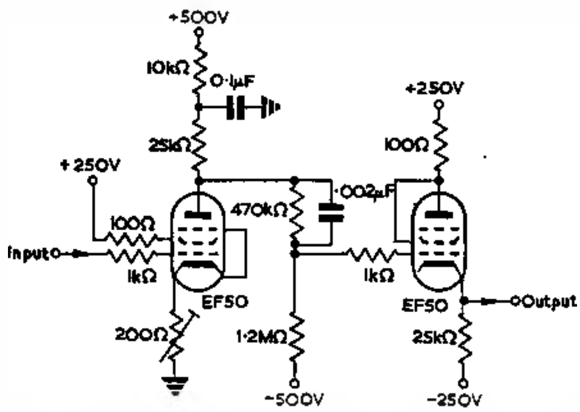


Fig. 2. Circuit of the standard computing amplifiers

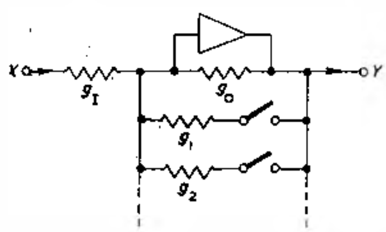


Fig. 3. Schematic circuit of the non-linear (squaring) unit

Non-Linear Units

Two main considerations influence the design of the non-linear units.

(1) The units should have a large bandwidth with negligible phase shift.

(2) The non-linear characteristics should be independent of changes in the characteristics of the individual non-linear elements—in this case germanium diodes.

These conditions may be fulfilled if the non-linear characteristic is generated by switching the feedback conductance of a feedback amplifier, as in Fig. 3, so as to alter the slope of the input-output characteristic in successive steps[5]. By making the number of switches which are closed in the feedback loop a function of the output (and hence of the input) voltage, a quadratic Y-X characteristic may be synthesized as a series of straight segments (see Fig. 4). This is equivalent to approximating the derivative of the quadratic characteristic (i.e., a linear graph) by a series of steps.

Fig. 5 shows the feedback circuit used to realize this scheme. R_o is a permanently connected feedback resistor and represents the maximum feedback resistance. $R_1 + R_1'$ is a feedback resistor which will be included in the feedback loop (in parallel with R_o) whenever the voltage at the junction of R_1 and R_1' due to the output is less than V_{B1} the bias voltage fed to the first diode D_1, by the low impedance bias chain formed by r_1 to r_{11}. When the output voltage of the non-linear unit rises above $V_{B1}(R_1+R_1')/R_1$ the diode D_1 conducts and clamps the junction point at V_{B1} so that the effective feedback resistance is decreased.

Thus the effective feedback resistance depends on the output voltage and hence the gain of the stage depends on the input voltage. By suitably choosing the resistors $(R_1 + R_1')$ to $(R_{12} + R_{12}')$ and the biasing points it can be arranged that the output voltage over a certain range is a quadratic function of the input (see Fig. 4).

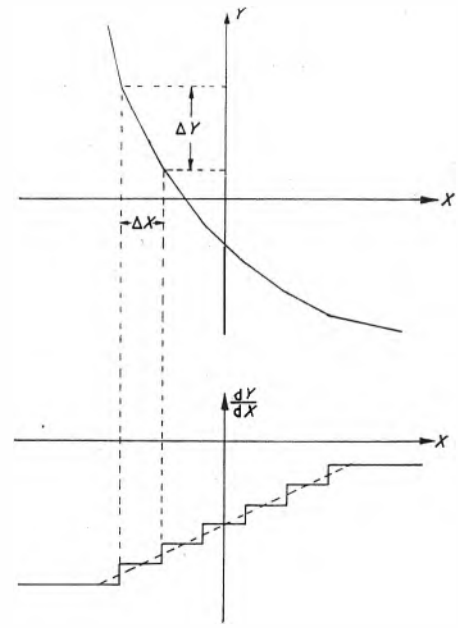


Fig. 4. Input-output characteristic of a non-linear unit, and the derivative characteristic showing the steps in gain

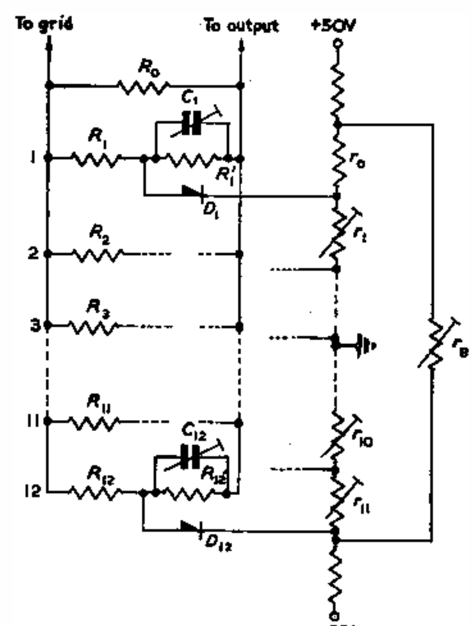


Fig. 5. Detailed circuit of the non-linear feedback network

The trimming capacitors C_1 to C_{12} across the feedback resistors R_1' to R_{12}' can be adjusted to compensate for the self-capacitances of the diodes. Thus (at least to the first order) the feedback is always non-reactive and a large bandwidth is possible.

The diode-switching occurs at a high signal level so that the feedback is substantially independent of secular changes in the diode parameters.

Choice of Parameters in the Non-Linear Unit

It can be seen from the circuit in Fig. 3 that the slope of the input-output characteristic cannot change sign: in other words a symmetrical parabolic characteristic cannot be generated. However, as shown already any portion of a parabolic characteristic may be used to obtain multiplication. The only disadvantage of an asymmetric characteristic is that a larger fraction of the input y must be subtracted to leave the product term xy. This in turn only means that the noise level of the output is slightly raised[9]. To minimize this effect, γ, the ratio of maximum slope of the characteristic, to its minimum slope should be chosen as large as practicable. A value of $\gamma = 5$ has been used in multipliers so far constructed. (Apart from considerations of convenience the possible magnitude of γ is limited practically by the gain of the amplifier and by the minimum permissible feedback resistance.)

The most important parameter of the non-linear unit is N, the number of steps in gain, i.e. the number of diodes. A theoretical analysis of the effect of the 'truncation' error (due to the finite steps in gain) which has been published elsewhere[9], shows that if the steps in gain are spaced at equal intervals along the input axis then the maximum error in the multiplier output, expressed as a fraction of the maximum undistorted output, is $1/N^2$. The root mean square error is about a third of this value. (The choice of equal intervals for the spacing of the steps in gain minimizes the mean square absolute error in the multiplier output when all values of the inputs x and y are equally probable. If small values of x and y are more probable, or if the errors for such inputs are more important, the spacing should be closer at the centre of the characteristic. For details reference 9 should be consulted.)

With 12 diodes the theoretical maximum error is 0·6 per cent of the maximum output. This figure was considered satisfactory for the proposed applications. An r.m.s. error of less than 0·01 per cent is obtainable with 20 diodes. As pointed out above the maximum value of the error will be attained for particular values of one input even when the other is zero so that the proportional error will be much greater than 0·6 per cent in unfavourable cases. For this reason the scale factors of the inputs to the multiplier must be chosen as large as possible.

Calculation of Feedback Resistors

Once N, the number of steps in gain, and γ, the gain ratio have been selected, suitable values for the resistors in the circuit of Fig. 5 may be calculated by using the fact that the gain of the non-linear unit (Fig. 3) with k switches closed is

$$\alpha_k = \frac{g_I}{g_o + \sum_{i=1}^{k} g_i}, \quad \alpha_o = g_I/g_o,$$

$$(k = 1, 2 \ldots N) \qquad (4)$$

If a distribution of equal steps in gain has been chosen then the feedback resistors are given by

$$R_k + R_k' = R_o(M - k)(M + 1 - k)/M \qquad (5)$$

where

$$M = N\gamma/(\gamma - 1) \qquad (6)$$

This formula neglects the finite resistance through the diode and the bias chain to earth which prevents the feedback through a given resistor from being cut off completely. However, if R_o and R_k' are large enough this is not serious. In the multipliers so far built R_o was taken as 200kΩ and the R_k' were 90kΩ or more each, compared with a maximum resistance through the bias chain of the order of 1kΩ. It is undesirable to choose very large values for the feedback resistors since this makes the required values of the trimmer capacitances too small to be mechanically stable.

The resistor values of the bias chain may be calculated from the relation

$$r_k/r_o = 1 - k/M \qquad (7)$$

This expression neglects the step down effect of the divider formed by R_k' and R_k. It is convenient to keep the ratio R_k'/R_k small. However, as the final values of the bias resistors are determined by practical adjustment this expression should only serve as a guide. A value of about 600Ω was used for r_o.

To minimize the resistance through the diodes to earth the junction of the bias chain which naturally lies nearest earth potential should be directly connected to earth. If this is the junction between r_m and r_{m+1} then m is roughly given by

$$m \simeq 0{\cdot}29M - 0{\cdot}9 \qquad (8)$$

The bias chain is optimally designed to utilize the full range of the output voltage of the amplifier; in the present case about ±15V. Varying the resistor r_B across the chain (Fig. 5) compresses the voltage over which the gain is stepped; this facility is useful for aligning the multiplier.

Construction

A set of three '12-diode multipliers' has been built. The photographs in Figs. 6 and 7 show one of these which fits side by side with the other two into a frame mounted on a post-office rack. Each multiplier may be slid out easily for adjustments.

The five amplifiers have been arranged on the chassis to give the shortest possible signal paths. The amplifiers are separately screened and coaxial cable is used for the longer signal paths. The two input terminals, the grid of the last stage and the output terminal are brought out on the steel front panel. (The grid terminal is provided so that the last stage can be converted to an integrator if required.)

The non-linear feedback circuits are mounted on a paxolin panel at the rear end of the chassis away from sources of heat (see photographs). GEX 54 germanium diodes were used for the 12-diode multipliers (a previously constructed nine-diode multiplier utilized CG6-Cs). These diodes have a large inverse resistance which makes them suitable for the present applications.

High stability 5 per cent tolerance resistors were used in all signal paths. The biasing chain was constructed from small wirewound potentiometers, which may be seen with the air trimming capacitors, on the left in Fig. 7.

The power consumption of a complete multiplier is about 40 mA at +500V, 50mA at +250V and −250V, 3mA at −500V, and 3A at 6·3V d.c. These figures are determined by the requirements of the standard amplifier. 9mA at +50V and −50V is used in the biasing chain of the non-linear units. It is obviously not essential to use such a large number of separate power supplies; with only minor modifications the complete unit could be made to run from ±250V supplies; but the voltages mentioned were conveniently available from the computer installation.

Alignment

The process of aligning the multiplier falls naturally into three stages:—

(a) Alignment of the non-linear feedback networks to ensure an accurate quadratic characteristic.

(b) Alignment of the multiplier as a whole to ensure cancellation of unwanted terms in the output.

(c) Elimination of phase shifts.

As a preliminary to these adjustments the inputs x and y are earthed and the bias levels of the five amplifiers adjusted (in succession) to make their outputs zero (the potentiometer P_1 being set at the centre of its range).

(a) Non-linear Network

The performance of the non-linear feedback circuits may be examined by supplying an accurately linear rising waveform to the x-terminal of the multiplier (y being earthed) and feeding the output of amplifier 2 (or 3) to a differentiating circuit—a standard computing amplifier with capacitive input and resistive feedback. (The derivative

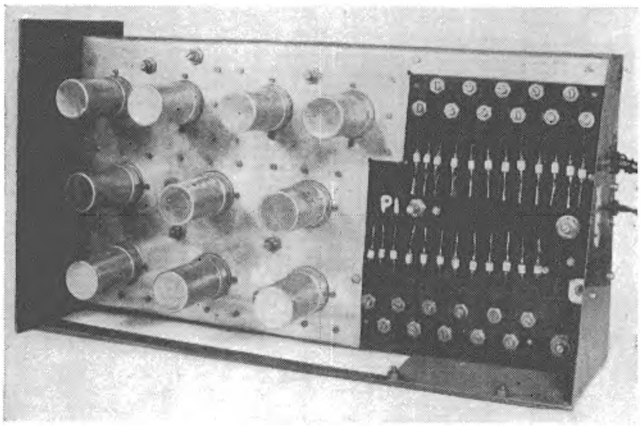


Fig. 6. View of a complete multiplier; the crystal diodes and the biasing potentiometers are on the paxolin panel

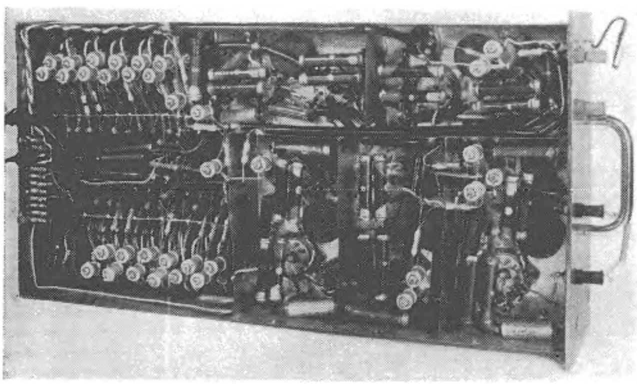


Fig. 7. Under-chassis view of a multiplier showing the screened amplifier compartments and the input and output terminals on the front panel

of the input should be examined to check that it is constant.) A typical display of the differentiated output versus the input is shown in Fig. 8; the stepped changes in the gain of the non-linear circuit are plainly visible. If the values of the feedback resistors are calculated by formula (5) without allowing for the forward resistance of the diode and bias chain then the steps of gain may be too small at the high gain end of the characteristic. This can be compensated for by increasing the value of the direct feedback resistor, R_0 in Fig. 5. In the present case the values were calculated on the assumption of $R_0 = 200\text{k}\Omega$ while $250\text{k}\Omega$ was found to be a suitable practical value.

If the non-linear characteristic is to be quadratic the graph of gain versus input must be linear. To check this a suitable fraction of the input may be subtracted from the differentiated output to give a voltage which should be constant. The resultant graph (versus input) is shown in Fig. 9; the steps in gain now appear as 'sawteeth'. By adjusting the potentiometers $r_1, r_2 \ldots r_{\text{ll}}$ in the bias chain each sawtooth can be made to intersect midway a reference line simultaneously displayed. Although the various adjustments are not completely independent the process is quite straightforward. The deviations of the sawteeth from the reference line indicate maximum errors in gain of 1·5 per cent (as a fraction of the maximum gain). Because of the curvature of the diode characteristics this is better than predicted by the theoretical formula

$$\Delta = 1/2M - (\gamma - 1)/2N\gamma \quad \ldots\ldots\ldots (9)$$

(where N is the number of steps in gain and γ is the gain ratio).

The two non-linear circuits in the multiplier are adjusted to be as similar as possible. When the overall bias resistor r_B in Fig. 5 is properly set both circuits will handle the same range of input voltages as may be verified from the display of gain versus input.

(b) Alignment of the Multiplier

When the non-linear networks are adjusted to give a quadratic characteristic, the output of the multiplier will in general be

$$z = Axy + Bx^2 + Cx + Dy^2 + Ey + F \quad \ldots\ldots (10)$$

neglecting the truncation errors due to the stepped gain. The object of aligning the multiplier is to make the coefficients

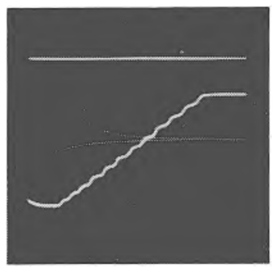


Fig. 8. Display of the differentiated output characteristic of a practical non-linear unit. (Compare with Fig. 4)

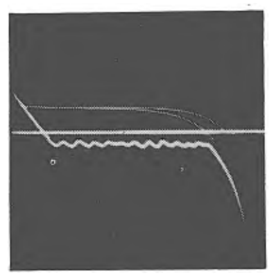


Fig. 9. Display of the differentiated output characteristic of a non-linear unit after subtraction of a fraction of the input

B, C, D, E and F zero; five controls should suffice for this. There is a wide latitude available in the choice of adjustable parameters. It is desirable practically that the adjustments may be performed in sequence and with little interaction. Theoretical analysis and practical experience have found the following scheme to be satisfactory.

The y-terminal of the multiplier is earthed ($y = 0$) and the characteristic output versus x-input displayed. Equation (10) shows that this will be a portion of a parabola. By adjusting K_1 (the feedback resistor of amplifier 1, see Fig. 1) the curvature may be eliminated (i.e. $B = 0$). The slope may now be reduced to zero ($C = 0$) by adjusting P_1 (which supplies the bias to the diode networks). The accuracy with which these adjustments can be carried out is limited only by the 'sawtooth' nature of the characteristic. (See Fig. 10 which shows the characteristic of output versus x-input after alignment).

The x-terminal is now earthed and the corresponding characteristic displayed. The curvature may be eliminated ($D = 0$) by adjusting K_4 (the feedback resistor of amplifier 4) and then the slope may be reduced to zero ($E = 0$) by adjusting P_2. The resulting characteristic is shown in Fig. 11.

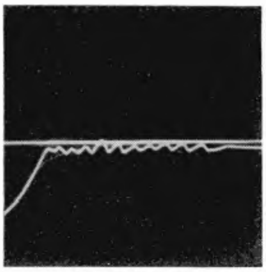
The d.c. component of the multiplier output (represented by F) is now eliminated by adjusting the bias on the final amplifier. The coefficient A may be made equal to unity (or any other value) by choosing a suitable value for the final feedback resistor, R_{out} in Fig. 1.

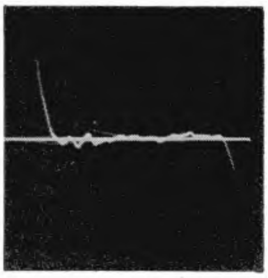
After the alignment the output of the multiplier will be the product, xy, plus the non-eliminable truncation errors (sawteeth). Figs. 10 and 11 show that these errors are less than 0·25 per cent of the maximum multiplier output whereas the theoretically predicted error is 0·6 per cent. The superior practical accuracy is due here, as in the case of the gradient errors, to the curvature of the diode characteristics.

(c) PHASE SHIFTS

Phase shifts occurring in the linear circuits of amplifiers 1, 4 and 5 are easily eliminated up to high frequencies (a 50kc/s square wave was used as test waveform) by trimming the 30pF air capacitors which are connected in parallel with each input and feedback resistor.

Owing to the self-capacitances of the germanium diodes, the phase shifts in the non-linear units are more serious. Fig. 12 shows the characteristic of input versus output for one non-linear circuit taken with a 50kc/s sine wave. No phase shift is visible with the trimmers properly adjusted. In practice there is a slight thickening of the cathode-ray





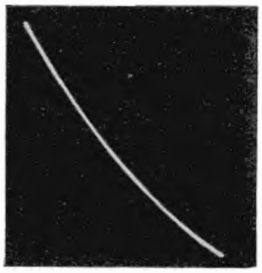


Fig. 10. Overall multiplier characteristic when the y-terminal is earthed, showing the nature of the zero error

Fig. 11. Overall multiplier characteristic with the x-terminal earthed

Fig. 12. Display of the input-output characteristic of a non-linear unit taken with a 50kc/s sinusoidal wave

tube trace which corresponds to a phase shift of 1° or less.

The adjustment of the trimmers in the non-linear circuit is somewhat complicated by interactions. It has been found that in these units no trimmer is needed across the direct feedback resistor, R_0 in Fig. 5, or across the first five feedback resistors at the high gain end of the ladder. Thereafter increasing values of capacitance are required up to a maximum of about 25pF across the resistor of lowest value.

The most critical test is to observe the multiplier output for a 50kc/s sine wave input to the terminal when the x-terminal is earthed. The phase shift errors in the two non-linear units then augment one another and the maximum zero error in the multiplier output increases to about $1\frac{1}{2}$ per cent.

STABILITY OF ALIGNMENT

The stability of the alignment is dependent primarily on three factors

(a) The stability of the various input and feedback resistors and (more importantly) of the biasing potentiometers.

(b) The constancy of the diode characteristic.

(c) The drifts of d.c. levels which in turn depend on the valves in the amplifiers and on the power supplies.

The last factor is the most serious, although vibration may alter the settings of some of the potentiometers, and (very rarely) a germanium diode may fail completely. Well stabilized h.t. supplies to the first stages of the amplifiers are essential. A 1 per cent change in h.t. voltage may produce an error in the multiplier output of 2 per cent. The supplies to the bias chain are equally important, the corresponding output error being slightly larger in proportion. To reduce hum, a direct current supply to the heaters is desirable. This must also be stabilized since a 1 per cent change in l.t. produces a 1·2 per cent output error.

Provided the multipliers have been allowed to warm up before use little trouble has arisen from changes in valve characteristics. None-the-less for slower speed applications drift compensation might be desirable. In practice the errors arising during a day or from one day to another have seldom exceeded 1 per cent, although it is advisable to check the alignment periodically.

Applications

As explained in the introduction the multipliers were intended for use in high speed repetitive analogue computing. An analogue computer equipped with a number of multipliers can handle many problems of interest besides

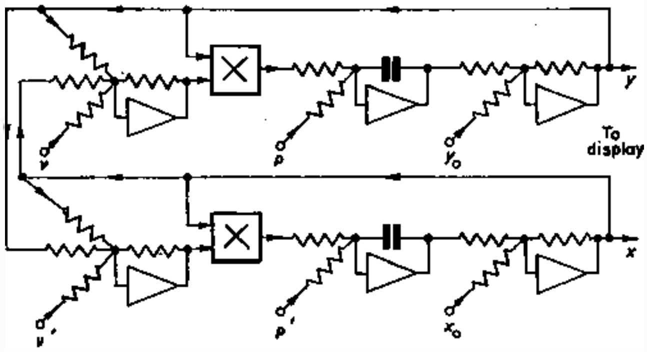


Fig. 13. Computer circuit for solving the non-linear 'stellar structure' equation

those of the usual linear differential equations. One such non-linear problem is the differential equation

$$dy/dx = \lambda \frac{y(x - \mu y + \nu) + \rho}{x(y - \mu' x + \nu') + \rho'} \quad \ldots\ldots\ldots (11)$$

A special instance of this equation ($\rho = \rho' = 0$) is the simplest equation of stellar structure described by Bondi[9].

If the usual analogue were adopted, with voltage representing the dependent variable y, and time the independent variable x, two multipliers and a divider would be needed to set up equation (11) on a computer. The requirement of a divider may be avoided[10], however, by introducing the time as a parametric variable and rewriting equation (11) in the form

$$dy/dt = \lambda[y(x - \mu y + \nu) + \rho] \quad \ldots\ldots\ldots (12)$$
$$dx/dt = x(y - \mu' x + \nu') + \rho'$$

It is evident that only two multipliers are required with this formulation of the problem. The appropriate circuit is shown in Fig. 13. The input voltages x_0 and y_0 determine the initial conditions; they fix the point in the $x-y$ plane from which the solution of equation (11) will run.

With a repetitive computer such as that for which the multipliers were designed the equations (12) may be solved 1 000 times a second (or more frequently). It is then useful to vary the initial conditions from one solution to the next so as to obtain a family of solutions in the $x-y$ plane instead of one single solution. In the computer at King's College[5] as many as 32×32 different conditions may be imposed cyclically. With an integrating period of 1msec for each solution a whole set of 32 solutions may be recomputed 30 times a second so as to give a steady display on a cathode-ray tube. Fig. 14 is a photograph of such a set of solutions to the non-linear equation (11). Figs. 15(a) and

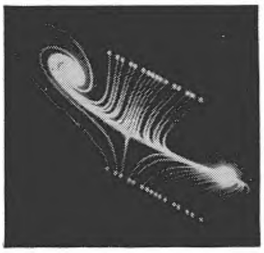


Fig. 14. Simultaneous display of a set of solutions of the 'stellar structure' equation

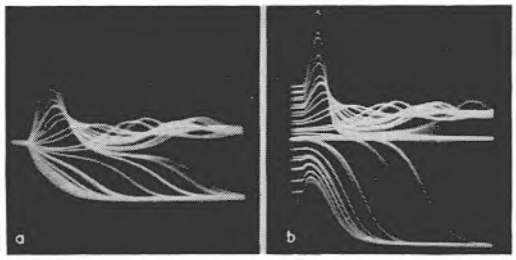


Fig. 15. Displays of sets of solutions versus the parametric variable:

(*a*) x *versus* t, (*b*) y *versus* t.

Fig. 17. Simultaneous display of a set of solutions obtained by automatically varying the parameter ν

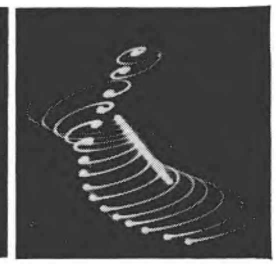


Fig. 18. A three-dimensional display of the set of solutions shown in Fig. 17 using the parameter ν **as the third co-ordinate**

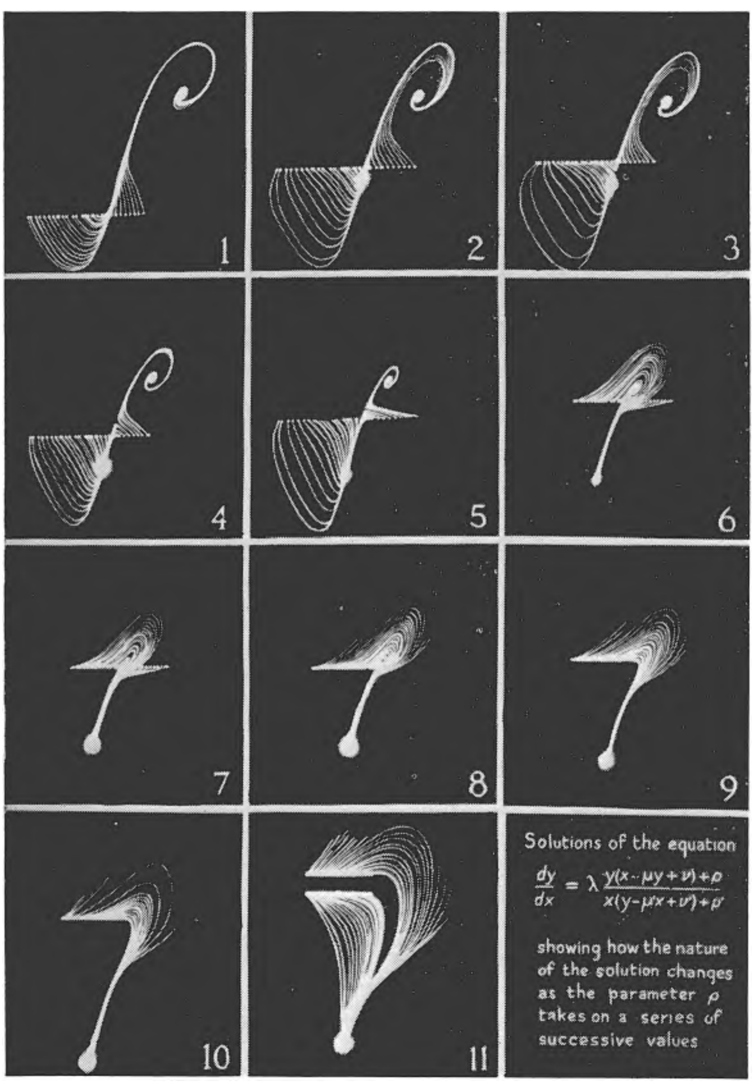
(b) show x and y plotted against t the parametric variable, for the same problem.

Since equation (11) is a first order differential equation there should be one and only one solution through each point in the $x-y$ plane; the solutions should never cross. That this is so can be seen from Fig. 14, the display of the solutions; at least there are only three points at which this generalization breaks down. These points (the centre of the spiral being one of them) are 'singular' points. They are points where the numerator and denominator of equation (11) vanish simultaneously.

In general there are four of these singular points, but for certain combinations of the parameters there may be fewer. Fig. 16 depicts a series of solutions of equation (11) for successive different values of the parameter ρ which was altered manually. It shows how the spiral grows smaller until in stage 7 it disappears after coalescing with the 'branching point' which first appears in stage 3.

Instead of altering a parameter manually to study the behaviour of the solution it may be changed automatically. Fig. 17 shows a set of solutions each with a slightly different value of the parameter ν. It will be seen that the centres of all the spirals lie on a portion of a hyperbola. Fig. 18 is a three-dimensional display[8] of the same set of solutions which separates out those with different values of the parameter.

Fig. 16. Solutions of equation (11)



In a problem as complex as equation (11) in which the individual solutions display a bewildering variety of forms, the ability to study sets of solutions and vary parameters automatically, is essential if insight is to be gained rapidly. This ease with which the nature of a problem can be resolved provides the justification for computing at the high repetition rates described.

Acknowledgments

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Control Circuit Design

By N. B. Acred*, Grad.Brit.I.R.E.

The article describes a diagram method for the simplification at Boolean algebraic expressions. Further development of the method introduces the concept of time sequence, which facilitates the direct design of control circuits.

WITH the advent of the 'automatic control era' new exercises in control engineering are created daily. In consequence, the electronic engineer is finding it necessary to apply his knowledge to the solving of problems which are in the new province of control-systems engineering. Usually system design is carried out by constructing a block diagram from the initial requirements, and then following the tedious task of tracing the route of the various signals from unit to unit within the diagram. To ease this task some form of organized approach is desirable, and the following is the description of a suitable method. The method does not necessarily arrive at the most economical design, in the circuit component sense, it does, however, present a reliable circuit design in a way which is easy to understand and simple to apply.

The basis for the design is Boolean algebra, an algebra of logic which is finding increasing usage among circuit design engineers. Since the fundamental concepts of the algebra are adequately presented elsewhere[1,2,3] it is not intended to reiterate them apart from the brief statements of Table 1.

When designing a circuit, one may express a particular desired effect by an equation which uses several circuit component functions. Since even in a simple system many such equations exist, severe difficulties may be experienced when trying to integrate them into a connexion scheme which is satisfactory for all combinations of the individual circuit variables.

Consider the case of four two-state variables, A, B, C and D; a total of 16 different combinations are available, as follows:

(1)	$\bar{A}\,\bar{B}\,\bar{C}\,\bar{D}$	(9)	$\bar{A}\,\bar{B}\,\bar{C}\,D$
(2)	$A\,\bar{B}\,\bar{C}\,\bar{D}$	(10)	$A\,\bar{B}\,\bar{C}\,D$
(3)	$\bar{A}\,B\,\bar{C}\,\bar{D}$	(11)	$\bar{A}\,B\,\bar{C}\,D$
(4)	$A\,B\,\bar{C}\,\bar{D}$	(12)	$A\,B\,\bar{C}\,D$
(5)	$\bar{A}\,\bar{B}\,C\,\bar{D}$	(13)	$\bar{A}\,\bar{B}\,C\,D$
(6)	$A\,\bar{B}\,C\,\bar{D}$	(14)	$A\,\bar{B}\,C\,D$
(7)	$\bar{A}\,B\,C\,\bar{D}$	(15)	$\bar{A}\,B\,C\,D$
(8)	$A\,B\,C\,\bar{D}$	(16)	$A\,B\,C\,D$

TABLE 1.
Boolean Identities.

1. Postulates

$+$ = OR		$.$ = AND
$A + 0 = A$		$A\,.\,0 = 0$
$A + 1 = 1$		$A\,.\,1 = 1$
$A + A = A$		$A\,.\,A = A$
	$\bar{A}$ = Not A	
$A + \bar{A} = 1$		$A\,.\,\bar{A} = 0$
	$\bar{\bar{A}} = A$	

2. Inversion Formulae

$$A + B + C = \bar{A}\,.\,\bar{B}\,.\,\bar{C} \qquad \overline{A\,.\,B\,.\,C} = \bar{A} + \bar{B} + \bar{C}$$

3. Reduction Formulae

$$A + \bar{A}\,B = A + B$$
$$\bar{A} + A\,B = \bar{A} + B$$
$$A\,B + A\,C + B\,.\,\bar{C} = A\,C + B\,\bar{C}$$

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Now suppose that any one of several combinations will produce a particular effect-function Z and that certain other combinations are inadmissible; possibly due to physical limitations or safety requirements, e.g. suppose that the combinations 5, 6, 13 and 14 must not exist, and the expression for Z is composed of the sum of the remaining 12 combinations.

$$\begin{aligned}\therefore Z = {} & \bar{A}\,\bar{B}\,\bar{C}\,\bar{D} + A\,\bar{B}\,\bar{C}\,\bar{D} + \bar{A}\,B\,\bar{C}\,\bar{D} + A\,B\,\bar{C}\,\bar{D} \\ & + \bar{A}\,B\,C\,\bar{D} + A\,B\,C\,\bar{D} + \bar{A}\,\bar{B}\,\bar{C}\,D + A\,\bar{B}\,\bar{C}\,D \\ & + \bar{A}\,B\,\bar{C}\,D + A\,B\,\bar{C}\,D + \bar{A}\,B\,C\,D + A\,B\,C\,D \qquad \ldots\ldots\ldots\ldots\ (1)\end{aligned}$$

This equation would normally be manipulated by the laws of Table 1 in an attempt to arrive at a simplified result. One of the difficulties of Boolean algebra is that there is no easy way of arriving at a simplified result other than by experience and/or inspiration. Another point to note is that during the manipulation of the equation each and every new equation will represent a physical system which is the equivalent of the original statement, and there is nothing to suggest that the simplest result is necessarily the most convenient one to use.

Diagram Method

A matrix construction method of representing the four variables has been suggested[3] and this forms the basis of the proposed system. By constructing a four by four matrix the 16 combinations of A, B, C and D can be inserted as follows.

	B	B	$\bar{B}$	$\bar{B}$	
A	4	12	10	2	$\bar{C}$
A	8	16	14	6	C
$\bar{A}$	7	15	13	5	C
$\bar{A}$	3	11	9	1	$\bar{C}$
	$\bar{D}$	D	D	$\bar{D}$	

The numbers inserted in the squares correspond to the 16 combinations of the four variables.

The explanation of the method of insertion is self evident; however, the following facts are of interest:

(1) The two combinations represented by numbers in any two adjacent squares differ by only one variable, e.g 10 and 12 differ by $\overline{B}$ and B.

(2) The same applies to squares at the end of rows or columns, e.g. 2 and 4 differ by $\overline{B}$ and B; 11 and 12 by $\overline{A}$ and A.

(3) Whole rows or columns; the combinations differ by two variables.

(4) Any four adjacent squares; the combinations differ by two variables, e.g. 4, 8, 12 and 16 differ by $\overline{C}$ and C, and $\overline{D}$ and D.

(5) The same applies to pairs of adjacent squares at the ends of adjacent rows or columns, e.g. 4, 8 and 2, 6, and to the four corner squares 1, 2, 3 and 4.

(6) Pairs of adjacent rows or columns; these combinations differ by three variables.

As an example of the utility of the above, consider the case where an effect Z is the result to the combinations represented by the two adjacent squares 4 and 12.

$$Z = A\,B\,\overline{C}\,\overline{D} \;+\; A\,B\,\overline{C}\,D \;=\; A\,B\,\overline{C}\,(\overline{D} + D) \;=\; A\,B\,\overline{C}$$

where $(\overline{D} + D) = 1$ (see Table 1)

Now by reference to the matrix it is seen that the combinations of squares 4 and 12 differ by $\overline{D}$ and D. Thus when considering a function which is the result of several combinations, and these combinations fall into the categories 1 to 6 referred to above, then a simplification of the variables may be effected by treating the differing variables as redundant.

Returning to the example of equation (1) and constructing a matrix, with 1 representing a required square and 0 an unrequired square.

	B	B	$\overline{B}$	$\overline{B}$	
A	1	1	1	1	$\overline{C}$
A	1	1	0	0	C
$\overline{A}$	1	1	0	0	C
$\overline{A}$	1	1	1	1	$\overline{C}$
	$\overline{D}$	D	D	$\overline{D}$	

Thus function $Z = B + \overline{C}$

where B is the two left-hand columns and $\overline{C}$ is the top and bottom rows.

This result, which is directly seen by inspection of the matrix, is less obvious from the equation (1).

A further variation is the case where there are free squares, these are squares which are neither specified by the initial equation as required, nor by other considerations as inadmissible. Any square which is not specified inadmissible must be admissible by definition, thus free squares are admissible.

Take an example using the same four variables A, B, C and D

Let $Z = A\,B\,\overline{C}\,\overline{D} + A\,B\,C\,\overline{D} + A\,B\,\overline{C}\,D + A\,B\,C\,D$

Also the following combinations are inadmissible,

$$\overline{A}\,\overline{B}\,\overline{C}\,\overline{D},\ \overline{A}\,\overline{B}\,C\,\overline{D},\ \overline{A}\,\overline{B}\,\overline{C}\,D,\ \overline{A}\,\overline{B}\,C\,D$$

Constructing the matrix and indicating the free squares by an X.

	B	B	$\overline{B}$	$\overline{B}$	
A	1	1	X	X	$\overline{C}$
A	1	1	X	X	C
$\overline{A}$	X	X	0	0	C
$\overline{A}$	X	X	0	0	$\overline{C}$
	$\overline{D}$	D	D	$\overline{D}$	

From the matrix $Z = A\,.\,B$ not using free squares

$$\left.\begin{array}{r}\text{but } Z = A\\ \text{or } Z = B\end{array}\right\}\ \text{using free squares.}$$

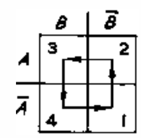
The second result is more convenient since it requires only one variable.

The method could be extended to include an extra variable, E, by cutting each square by a diagonal and calling the upper and lower sections the two possible states of the new variable. For more than five variables the method is impracticable; however, the complete control system can be sub-divided into up to five sections and then each section further sub-divided as necessary. This limitation of the number of variables which can be handled simultaneously corresponds to the limitation in variable economy.

System Design

A control system is specified by the number, type and order in which its outputs should be available, according to specific input instructions. Given a problem, the engineer must first decide how he is going to group and sub-group his dependent variables, keeping in mind the order in which the various operations are to be carried out. He will then draw up a table showing the sequence in which the variables change their states; from this table the matrix diagram may be directly constructed. Consider a simple sequence and its diagram.

	Variables A	Variables B
Sequence No. 1	0	0
2	1	0
3	1	1
4	0	1
1	0	0
	etc.	



Arrows show route of sequence within the diagram.

From the diagram the designer can establish the operate, hold and release conditions (to use relay terms) for each of the variables. In the case of a bistable device, only the operate and release conditions are of interest, because the hold condition is an inherent feature. With reference to the above diagram, square 1 corresponds to the operate, square 2 to the hold, and square 3 to the release conditions of the variable A.

A new variable has now been introduced, i.e. the sequence number or operation time-period. As a sequence number it will show the order in which the operations will naturally occur due to the circuit interconnexions, or as a time-period it will show the particular time at which an operation will occur due to a combination of circuit conditions and a reference pulse timing signal.

When designing a system it is necessary to arrange for a change in one variable at a time, this results in a continuous and non-ambiguous route within the diagram which, of course, implies non-ambiguous operation.

The following are three design studies using the diagram method.

Example No. 1

Consider an automatic drilling operation, consisting of a conveyor belt, an automatic positioning device, and a drill. The conveyor is halted when an item is beneath the drill, a pneumatically operated arm accurately positions the workpiece, and the drilling to a fixed depth follows. After drilling the item is released and the conveyor continues.

The primary sequence is first considered using the following variables.

V = the conveyor belt.

P = the pneumatic positioning device.

D = the drilling operation.

X = the proximity switch indicating that the item is approximately in the drilling position.

Sn = sequence number.

Sn	V	P	D	X	Y	
0	0	0	0	0	0	off condition
1	1	0	0	0	0	conveyor belt moving
2	1	0	0	1	0	proximity switch operates
3	0	0	0	1	0	conveyor stops
4	0	1	0	1	0	pneumatic positioning
5	0	1	1	1	0	drilling commences
6	0	1	1	1	1	
7	0	1	0	1	1	drilling ends
8	0	0	0	1	1	workpiece released
9	1	0	0	1	1	conveyor proceeds
10	1	0	0	0	1	
1	1	0	0	0	0	

The other variable Y has been introduced to differentiate between sequence numbers 1 through 5 and 6 through 10 and thus provide a non-ambiguous route within the sequence diagram. This extra variable can be used in the drilling sequence.

The sequence diagram follows:—

	P		$\bar{P}$		P		$\bar{P}$		
V	X	X	9	10	X	X	2	1	$\bar{D}$
	X	X	X	X	X	X	X	X	D
$\bar{V}$	X	6	X	X	X	5	X	X	
	X	7	8	X	X	4	3	0	$\bar{D}$
	$\bar{X}$	X		$\bar{X}$		X		$\bar{X}$	
	Y				$\bar{Y}$				

The system equations may be taken directly from the diagram. To illustrate the method this will be done in greater detail than is necessary.

Variable	Operate Conditions	Hold Conditions	Equation
P	S_3	$S_{4,5,6}$	$D + \bar{Y} . \bar{V} . X$
D	S_4	$S_{5,6}$	$D + \bar{Y} . P$
Y	S_5	$S_{6,7,8,9}$	$D + Y . X$
V	S_8	$S_{9,10}$	$Y . \bar{P}$
V	S_0	S_1	$\bar{X} . START \cdot \overline{STOP}$

When extracting the equations, always include at least one sequence square in two terms; this obviates the possibility of release during switching transitions; for example $D = D + \bar{Y} P$; the sequence square S_5 is included in both terms of the equation.

The release of variable D is not wholly dependent upon the primary variables and occurs when the drilling sequence is completed, thus the sequence square S_6 is considered to be a hold condition.

Conveyor Operation

The variables associated with the conveyor are:—

V; the conveyor power source.

X; the proximity switch.

and the start/stop switch.

$$V = Y . \bar{P} + \bar{X} . START . \overline{STOP}.$$

Positioning Device

The positioning is accomplished by a solenoid-operated two-way valve which is fitted with back-pressure contacts in either direction. Let the operate back-pressure switch represent contact P, and the release back-pressure switch represent contact $\bar{P}$.

Drilling Sequence

D = drill spindle power

F = drill feed power

Y = feed direction

L_1 = upper position limit switch

L_2 = lower or drilling depth, limit switch

The required operational table and sequence diagram is as follows. The numbers within the diagram represent the sub-sequence Qn.

Sn	Qn	D	L_1	L_2	Y	F
4	0	0	1	0	0	0
5	1	1	1	0	0	0
	2	1	1	0	0	1
	3	1	0	0	0	1
	4	1	0	1	0	1
	5	1	0	1	0	0
	6	1	0	1	1	0
	7	1	0	1	1	1
	8	1	0	0	1	1
	9	1	1	0	1	1
6	10	1	1	0	1	0
7	11	0	1	0	1	0

	L_1		$\bar{L}_1$		L_1		$\bar{L}_1$		
F	X	X	X	X	X	X	X	X	$\bar{D}$
	9	X	7	8	2	X	4	3	D
$\bar{F}$	10	X	6	X	1	X	5	X	
	11	X	X	X	0	X	X	X	$\bar{D}$
	$\bar{L}_2$	L_2		$\bar{L}_2$		L_2		$\bar{L}_2$	
	Y				$\bar{Y}$				

and the equations $F = D (\bar{Y} . \bar{L}_2 + Y . \bar{L}_1)$

$$Y = \bar{F} . L_2 + Y$$

$$D = F + \bar{Y} + \bar{L}_1$$

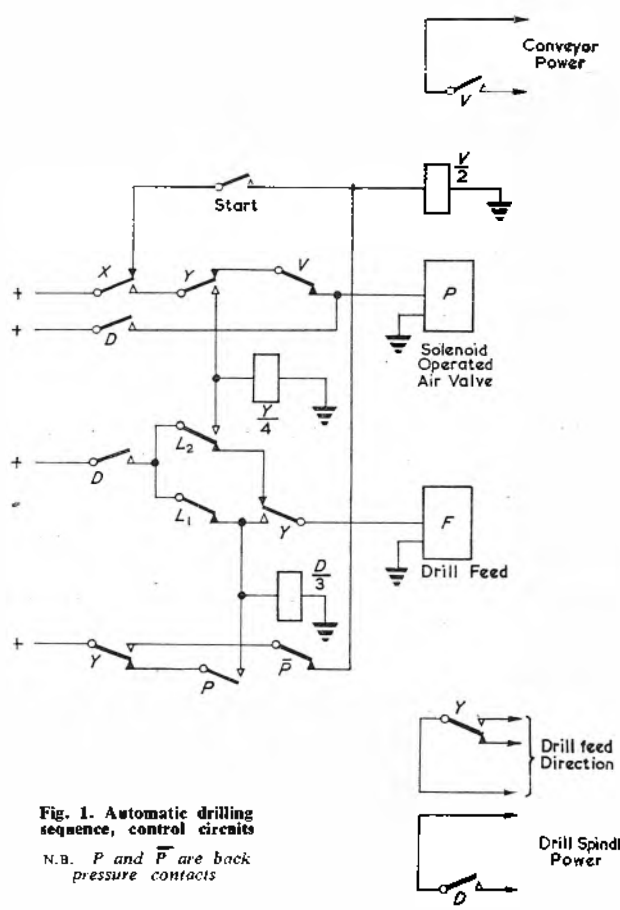


Fig. 1. Automatic drilling sequence, control circuits

N.B. *P and $\bar{P}$ are back pressure contacts*

Since F is not a function of itself, it may be eliminated from the equations. Inverting F according to table 1.

$$\overline{F} = \overline{D} + (Y + L_2)(\overline{Y} + L_1) = \overline{D} + (Y\,L_1 + \overline{Y}\,L_2 + L_1\,L_2)$$

inserting $\overline{F}$ into the equation for Y gives

$$Y = L_2 + Y$$

also

$$D = \overline{Y} + \overline{L_1}$$

To obtain the operational equations for the variables Y and D, the above results must be combined with those of the primary equations.

Thus $Y = (L_2 + Y)(D + Y \,.\, X) = D \,.\, L_2 + D \,.\, Y + Y \,.\, X$

The first term is the operate condition, the second term the hold condition due to the sub-sequence, and the third term the hold condition due to the primary sequence. In practice the second term is a redundant by-product of the combination of the two initial equations, and can be discarded.

also $D = (\overline{Y} + \overline{L_1})(D + \overline{Y} \,.\, P) = \overline{Y} \,.\, P + D \,.\, \overline{Y} + D \,.\, \overline{L_1}$

here again the second term is not strictly necessary.

Interpreting Y as a make contact and $\overline{Y}$ as a break contact, the circuit diagram for the control system may be drawn up from the foregoing equations; Fig. 1.

It must be emphasized that this example is intended to demonstrate the use of the matrix diagram method as a design tool, and is not tendered as a practical design.

Example No. 2

A translator is required to convert the reflected binary-coded output of a shaft digitizer, into true binary-code prior to computation. The two codes are presented below with their equivalent decimal values.

Reflected Binary				Decimal Value	True Binary			
d	c	b	a		D	C	B	A
0	0	0	0	0	0	0	0	0
0	0	0	1	1	0	0	0	1
0	0	1	1	2	0	0	1	0
0	0	1	0	3	0	0	1	1
0	1	1	0	4	0	1	0	0
0	1	1	1	5	0	1	0	1
0	1	0	1	6	0	1	1	0
0	1	0	0	7	0	1	1	1
1	1	0	0	8	1	0	0	0
1	1	0	1	9	1	0	0	1
1	1	1	1	10	1	0	1	0
1	1	1	0	11	1	0	1	1
1	0	1	0	12	1	1	0	0
1	0	1	1	13	1	1	0	1
1	0	0	1	14	1	1	1	0
1	0	0	0	15	1	1	1	1

Transferring to the matrix diagrams, and inserting the decimal values of the combinations into the squares.

	b		$\overline{b}$		
a	2	13	14	1	$\overline{c}$
a	5	10	9	6	c
$\overline{a}$	4	11	8	7	c
$\overline{a}$	3	12	15	0	$\overline{c}$
	$\overline{d}$	d	d	$\overline{d}$	

DIAGRAM R

	B		$\overline{B}$		
A	3	11	9	1	$\overline{C}$
A	7	15	13	5	C
$\overline{A}$	6	14	12	4	C
$\overline{A}$	2	10	8	0	$\overline{C}$
	$\overline{D}$	D	D	$\overline{D}$	

DIAGRAM T

The virtue of the reflected binary code is that between adjacent decimal values there is a change in only one of the four variables, which eliminates gross ambiguity in shaft position. This non-ambiguity is represented on the diagram by the fact that adjacent decimal numbers occupy adjacent squares, or squares at the ends of the same row or column.

Each of the true binary variables will be expressed in terms of the reflected binary variables.

e.g. True binary A is represented by squares 3, 7, 11, 15, 9, 13, 1, 5 of diagram T, and is equivalent to the sum of the combinations in reflected binary variables obtained from the same numbered squares in diagram R.

$$A = a\,b(c\,\overline{d} + \overline{c}\,d) + a\,\overline{b}(c\,d + \overline{c}\,\overline{d}) + \overline{a}\,b(\overline{c}\,\overline{d} + c\,d) + \overline{a}\,\overline{b}(\overline{c}\,d + c\overline{d})$$

$$\overline{A} = a\,b(\overline{c}\,\overline{d} + c\,d) + a\,\overline{b}(\overline{c}\,d + c\,\overline{d}) + \overline{a}\,b(c\,\overline{d} + \overline{c}\,d) + \overline{a}\,\overline{b}(c\,d + \overline{cd})$$

$$B = b(\overline{c}\,\overline{d} + c\,d) + \overline{b}(\overline{c}\,d + c\,\overline{d})$$

$$\overline{B} = b(c\,\overline{d} + \overline{c}\,d) + \overline{b}(c\,d + \overline{c}\,\overline{d})$$

$$C = c(\overline{d}) + \overline{c}(d)$$

$$\overline{C} = \overline{c}(\overline{d}) + c(d)$$

$$D = d$$

$$\overline{D} = \overline{d}$$

Inserting true binary variables into the equation, yields;

$$A = a\,\overline{B} + a\,\overline{B} \qquad B = b\,\overline{C} + \overline{b}\,C$$

$$\overline{A} = a\,B + \overline{a}\,\overline{B} \qquad \overline{B} = b\,C + \overline{b}\,\overline{C}$$

$$C = c\,\overline{D} + \overline{c}\,D \qquad D = d$$

$$\overline{C} = c\,D + \overline{c}\,\overline{D} \qquad \overline{D} = \overline{d}$$

The form of the above equations suggests the general expression

$$An = an \,.\, \overline{An_{+1}} + \overline{an} \,.\, An_{+1}$$

$$\overline{An} = an \,.\, An_{+1} + \overline{an} \,.\, \overline{An_{+1}}$$

which exhibit the same logic as a conventional binary half-adder.

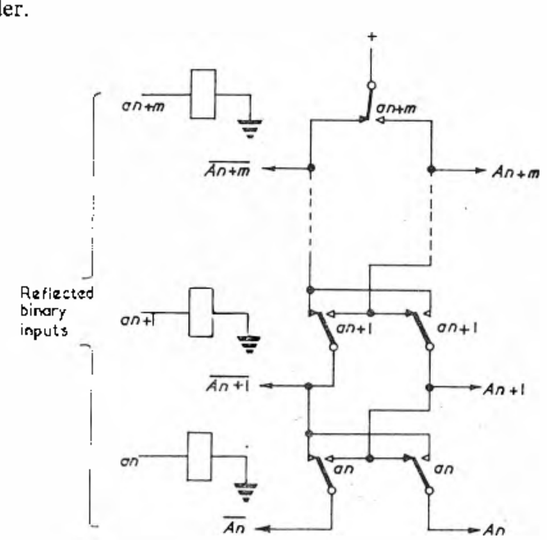


Fig. 2. Reflected-binary to true-binary convertor

A simple solution to the problem is obtained by using relays as the logical elements; the circuit is shown in Fig. 2.

Using the same method to derive the true binary to reflected code translation, the following general expressions are easily derived from the diagrams T and R.

$$an = An \,.\, \overline{An_{+1}} + \overline{An} \,.\, An_{+1}$$

$$an = An \,.\, An_{+1} + \overline{An} \,.\, \overline{An_{+1}}$$

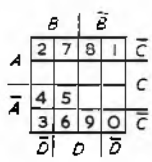
Example No. 3

It is required to design a diode interconnexion matrix which will translate a four 'bit' reflected-binary coded decimal system into a one-out-of-ten decimal system.

The binary code is as follows:

Decimal Value	Reflected Binary Value			
	D	*C*	*B*	*A*
0	0	0	0	0
1	0	0	0	1
2	0	0	1	1
3	0·	0	1	0
4	0	1	1	0
5	1	1	1	0
6	1	0	1	0
7	1	0	1	1
8	1	0	0	1
9	1	0	0	0

Transferring to the diagram and inserting the decimal values of the combinations in the squares.



The interpretation of the diagram is different in this case, since each output corresponds to only one combination of the four input variables. Because it is required to differentiate between individual outputs, it is the differing variables between adjacent squares which are of value. Thus the necessary combinations of the input variables for each of the ten outputs is easily derived from the diagram by stating the differing variables.

Decimal Value	Necessary Input Variables			
0	$\bar{D}$		$\bar{B}$	$\bar{A}$
1	$\bar{D}$		$\bar{B}$	A
2	$\bar{D}$		B	A
3	$\bar{D}$	$\bar{C}$	B	$\bar{A}$
4	$\bar{D}$	C		
5	D	C		
6	D	$\bar{C}$	B	$\bar{A}$
7	D		B	A

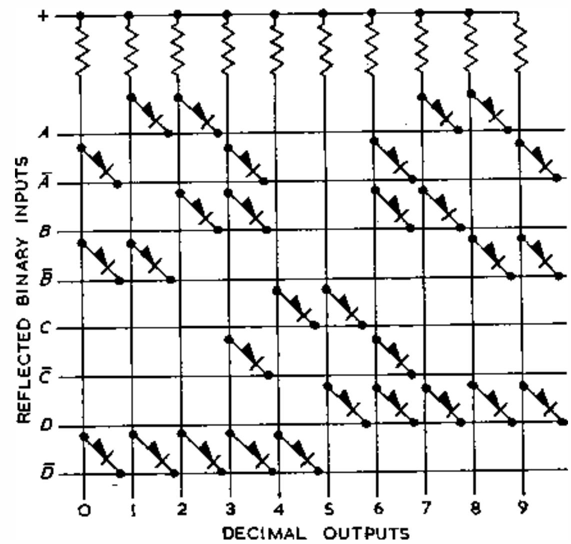


Fig. 3. Reflected-binary to decimal, code-translator

An input or output signal is indicated by a positive voltage

Decimal Value	Necessary Input Variables		
8	D	$\bar{B}$	A
9	D	$\bar{B}$	$\bar{A}$

The resulting diode matrix is shown in Fig. 3 and requires a total of 30 diodes. It can be seen that the organized approach has reduced mental fatigue and uncertainty to a minimum.

Conclusions

The matrix diagram method, as outlined in the previous paragraphs, is being currently used in the circuit design of industrial control equipment. The method contributes towards a better understanding of the operational sequence, and enables engineers, relatively inexperienced in this field, to produce effective and reliable designs.

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Computer Standardization

Simplifying the field of computer equipment through standardization is one of the aims of the Data Processing Section of the Radio Communication and Electronic Engineering Association which is now meeting regularly under the chairmanship of Mr. C. Metcalfe (EMI Electronics Ltd).

Four working parties set up by the Data Processing Section's techni al sub-committee have already gone a considerable way towards the formulation of standards. They are working closely with the British Standards Institution.

One working party has just completed the task of drawing up a chart for work on the many components and nomenclature used in computer fields.

Another party is considering standards for tape and will have a specification ready to submit to the B.S.I. before the end of this year. This will cover the requirements of both analogue and digital computers.

The production of specifications for an input keyboard and for a simple output device is being dealt with by a third working party. Yet another group is now studying information storage cores.

On the main committee one of the matters of joint interest receiving attention is the question of the many associations and sections building up in the United Kingdom to discuss data processing equipment and it is the aim of the R.C.E.E.A. Data Processing Section to see if a joint organization can be formed through which could be pooled all matters of national interest.

As already announced, the Radio Communication and Electronic Engineering Association is collaborating with the Office Appliance and Business Equipment Trades Association in organizing the first British Electronic Computer Exhibition to be held at Olympia, London, from November 28 to December 4, 1958. The Data Processing Section of R.C.E.E.A. is concerned in this and in the arrangements for the symposia, technical and non-technical, to be held in connexion with the Exhibition.

Applications of a

New Type of Cold Cathode Trigger Tube

(Part 3)

By G. O. Crowther*, B.Sc., A.C.G.I., and K. F. Gimson*, B.Sc.Eng., A.M.I.E.E.

Electronic Counting and Switching Circuits

This is a wide and rapidly growing field covering such applications as scientific computing, commercial computing, industrial control, batch counting, automatic telephone switching, nucleonic scaling, etc.

Although the basic principles are common throughout the range the individual techniques used depend very much on the speed and complexity of the task in hand and this in turn will dictate the cases in which the cold-cathode tube provides the best solution to a particular problem. Thus in most large scientific computers the operating speeds of the arithmetic units are above the range of cold-cathode tubes. The input and output systems, however, are relatively slow and cold-cathode tubes will work in conjunction with the input or card reader and also operate relays for punching the output. In slow speed commercial and industrial machines it is possible to use cold-cathode tubes throughout, i.e. in switching, memory, relay operation.

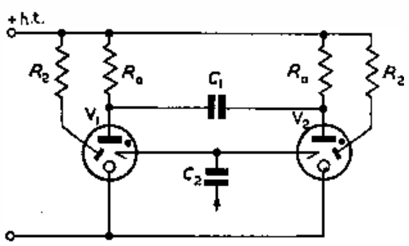

Fig. 35. Basic binary circuit

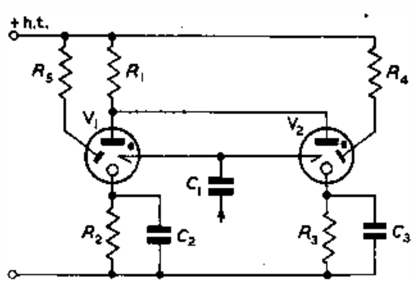

Fig. 36. Binary circuit with common anode load

In the allied application of telephone switching cold-cathode tubes may be used for relay, counting memory work and also for direct speech passing. The latter subject is beyond the scope of this article and the present section is devoted purely to a consideration of basic electronic switching circuits.

There are a number of well-known binary and ring circuits whose design is considerably simplified by the use of tubes with stable characteristics and a small tube to tube spread. Figs. 35 and 36 show two basic forms of binary circuits and Fig. 37 shows the basic binary circuit of Fig. 36 converted into a ring of n counter.

It should be noted at this point that the Z803U can only be used in the basic binary counters, not in the ring counter circuit. The reason for this is that the trigger of the conducting tube in the ring circuit is taken to cathode potential via R_2, R_3, and therefore it will act as if it is a cathode. It has already been stated that stability of the tube will be adversely affected if the tube is run under these conditions.

For the above reasons and because it is felt that the cold-cathode ring counter gives a simpler circuit than the ring of trigger tubes, the remarks made here will be mainly concerned with the two binary circuits. However, it should be noted that the ring circuits could be of advantage when printed circuit techniques and small subminiature trigger tubes are employed to give a more compact assembly. There will also be uses for these ring circuits in systems working in scales other than binary or decade.

Circuit Details

The operation of the two binary circuits is briefly as follows. Assume for the present that in Fig. 35 V_1 is conducting and therefore V_2 is non-conducting. The anode potential of V_1 is equal to V_m and that of V_2 is at h.t. The potential across C_1 is thus $V_{ht} - V_m$.

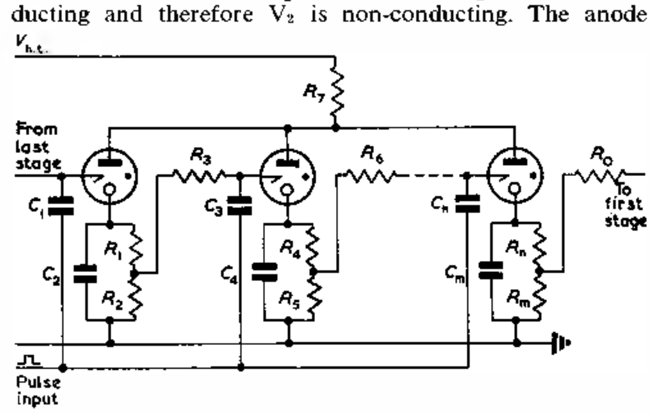

Fig. 37. Typical trigger tube ring counter

On the receipt of a pulse on the common trigger line V_2 is triggered and its anode potential falls to V_m. Because of the coupling capacitor C_1 the anode potential of V_1 will at the same instant fall by an equal amount to $(2V_m - V_{ht})$ volts and thus extinguish V_1. In general the anode potential will be below the cathode potential since V_{ht} is normally greater than $2V_m \simeq 210$V. The anode potential of V_1 is returned exponentially to h.t. with a time-constant

$$C_1\left(R_a + \frac{R_t R_a}{R_a + R_t}\right) \simeq CR_a$$

Where R_t is the internal impedance of the conducting tube V_2, and is normally negligible.

Once the anode potential of V_1 has been returned substantially to h.t. potential the circuit is ready to receive the next pulse on the common trigger line.

The circuit of Fig. 36 operates in a somewhat different manner. Assume again that V_1 is conducting and that the cathode potential of V_1 is V_k and the common anode potential is V_a. The following equations can then be written.

$$V_a = V_m + V_k$$
$$V_{ht} = iR_1 + V_a$$
$$V_k = iR_2$$

* *Mullard Ltd.*

The value for V_{ht} can be greater than $V_{a(max)}$ and preferably as high as is economically possible. The anode-cathode potential of V_2 is equal to V_a and to ensure that V_2 only conducts when a pulse is applied to the trigger $V_{a(min)} < V_a < V_{a(max)}$, where $V_{a(min)}$ and $V_{a(max)}$ are the limits of the anode working voltage range.

If now a pulse is applied to the common trigger line V_2 is triggered. Momentarily its cathode potential is held at earth by the capacitor C_3 and thus the potential of the common anode line falls to V_m. Similarly, due to the presence of C_2 the cathode potential of V_1 remains momentarily at V_k and thus the anode-cathode voltage of V_1 is reduced by $(V_a - V_m)$ volts and the tube is extinguished. The cathode potential of V_2 is then raised exponentially to V_k with time-constant.

$$\tau_1 = C_3 \frac{R_1 R_3}{R_1 + R_3} \text{ (neglecting } R_1\text{)}$$

The cathode potential of V_1 is returned exponentially to earth with a time-constant $\tau_2 = R_2 C_2$. The anode-cathode

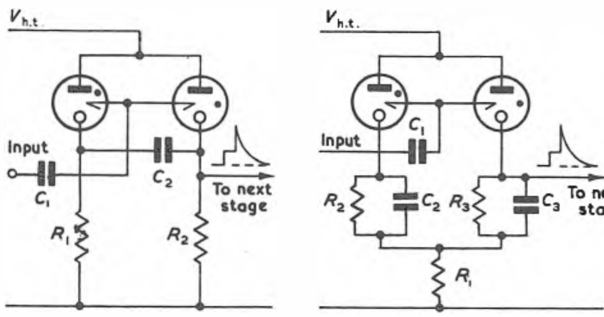

Fig. 38 (left and 39 (right) Inverted binary circuits

potential of V_1 is, therefore, increased to V_a according to the expression,

$$V_{1ak} = V_a - V_k \exp(-t/\tau_1) - V_k \exp(-t/\tau_2)$$

The two time-constants are selected to give the tube sufficient time to deionize.

The two binary circuits described have three completely stable states. Two are the conventional states with one tube conducting and the other non-conducting as already discussed. The third is with both tubes conducting. The last state is completely stable and to change to one of the other two states the supply voltage must be momentarily removed. With careful design the circuits can be made such that the third state is extremely difficult to establish.

The circuit shown in Fig. 35 has a number of advantages over that shown in Fig. 36. It can be designed as a bistable, monostable or an astable circuit whereas the circuit of Fig. 36 can only be readily designed as a bistable circuit. It can be arranged to give an output pulse which is more suitable for driving a secondary binary stage than that obtained when using the circuit of Fig. 36. It uses fewer components.

The second advantage is only strictly true when tubes such as the Z803U are employed. If a tube is employed in which the trigger can be connected to cathode during construction, the two inverted circuits shown in Figs. 38 and 39 give outputs suitable for driving another binary pair.

Figs 40, 41 and 42 show respectively the astable monostable and bistable versions of the binary circuit of Fig. 35. In the circuit of Fig. 40 assume that V_1 is conducting and V_2 non-conducting. As explained earlier, provided the current through R_1 is small the trigger will act as a probe in the main discharge and the potential of the trigger will be just below the burning voltage. The potential across capacitor C_2 will initially be less than $V_{t(ign)}$ but will be charged via the resistor R_2 towards h.t. potential. When it has charged to $V_{t(ign)}$ of V_2, V_2 will be triggered and C_2 partly discharged to a potential of just below V_m. V_1 will extinguish as in the circuit of Fig. 35. The trigger of V_1 is now no longer held by the discharge and the capacitor C_1 is charged via R_1 towards h.t. potential. Meanwhile, the trigger of V_2 is held just below the burning potential by the main discharge. Thus the circuit will 'flip-flop' at a rate determined by the components R_1C_1, R_2C_2, the burning voltage, the trigger ignition potentials of the tubes and the h.t. potential. The two tube characteristics are stable and thus the period of oscillation can be made stable.

It is essential that the anode of the non-conducting tube is returned to h.t. potential prior to trigger re-ignition.

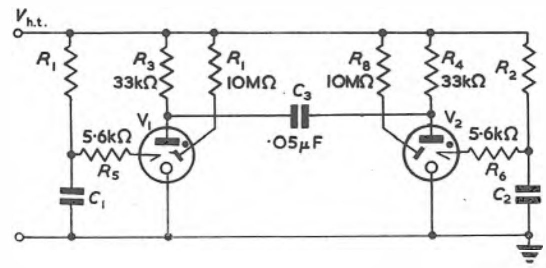

Fig. 40. Astable binary circuit

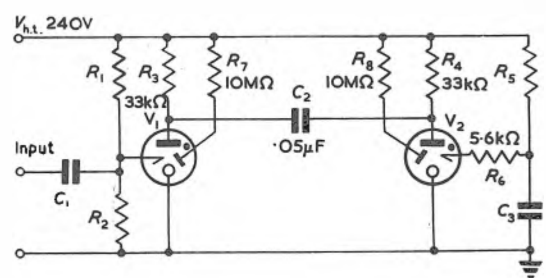

Fig. 41. Monostable binary circuit

Unless this is ensured one of two things may happen.

(1) The trigger discharge may fail to transfer to the anode, in which case the trigger capacitor must be recharged to $V_{t(ign)}$ and the circuit will remain twice as long in the particular state.

(2) The tube in question may be triggered, but because of the small fall in anode potential it will fail to extinguish the second tube. In this case both tubes will remain conducting and the circuit cease to operate.

For the Z803U, it is therefore recommended that a minimum safe anode time-constant of 2msec is used at all times. The anode resistors are already determined by the required anode currents through the two tubes and therefore capacitor C_3 is selected to give the correct time-constant. If for some reason either tube is required to pass a current in excess of 15 to 20mA, the value of the anode time-constant T_1 must be increased above the minimum value because of the rise in recovery time illustrated in Fig. 6. It should also be noted that if a relay forms a part of the anode resistance of the tubes the inductance will cause a more rapid rise in the potential than would occur with a pure resistance and again, the value of T_1 must be increased.

The maximum frequency at which a circuit of this form can be made to oscillate is 300c/s. The minimum frequency is, as in the timer circuit, a question of economics. It is probable that a frequency of 1c/min is the minimum.

The considerations in the design of the trigger circuit are identical to those of the timer circuit. The stability can be estimated using the formula in which the capacitor is charged from an initial voltage $V_o \simeq V_m$. To improve the stability of the circuit the trigger capacitors can be charged from a stable reference voltage rather than from the h.t.

In the circuit of Fig. 41, a monostable circuit is shown in which the stable state is with V_1 non-conducting and V_2 conducting. The trigger potential of V_1 is set below $V_{t(ign)}$ by the potential divider R_1 and R_2. The trigger of V_2, the conducting tube, is as before, held just below V_m by the discharge. On receipt of a positive pulse on the trigger of V_1, V_1 is triggered and V_2 extinguished. The circuit resets to its stable state after the period in which capacitor C_3 is charged from V_m to $V_{t(ign)}$. Except for the trigger circuit of V_1 the general considerations in the design of the circuit are identical to those previously described for the astable circuit. The trigger circuit of V_1 must be designed to ensure that the trigger discharge is automatically extinguished when the circuit resets back to its stable state. The simplest method is to select R_1 and R_2 to give a potential less than V_m. There are then no restrictions on the actual values of R_1 and R_2, but the circuit is comparatively insensitive. To obtain a more sensitive circuit the trigger potential can be set just below $V_{t(ign)}$. However, to make the circuit self extinguishing, the effective trigger resistance $R_1R_2/(R_1 + R_2)$ must now be greater than 1MΩ.

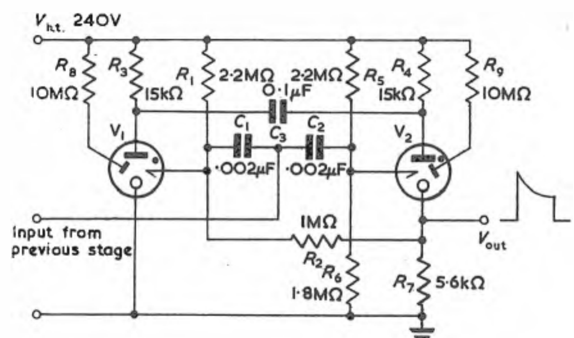

Fig. 42. Bistable binary circuit

Unless precautions are taken the basic bistable circuit of Fig. 35 has the disadvantage that during the switching on period it can be triggered easily into the state in which both tubes are conducting. When h.t. is applied to the circuit neither tube will conduct until a pulse is applied on the common input line to both triggers. The anode potentials of both the tubes will be at full h.t. potential and provided there is not a considerable delay in one of the tubes firing, both V_1 and V_2 will trigger. The anode potential of both tubes will therefore fall simultaneously by approximately the same amount to V_m and since the potential across the capacitor C_1 was initially zero, neither tube will be extinguished. The slight differences in V_m due to tube to tube spread and therefore in the fall of anode potential, will not be sufficient to extinguish the tube with the higher V_m. This is because the potential across C_2 will have been readjusted well before the tube has deionized. This trouble may be overcome by breaking the circuit of one tube until the other is conducting. It should be noted that once one of the tubes is conducting it is extremely difficult to establish conduction in the other tube without extinguishing the first.

An alternative method of avoiding this trouble is shown in Fig. 42. The trigger potential of V_1 is made less than that of V_2 and the potentiometer chain for the trigger of V_1 is taken to the output resistor in the cathode of V_2 instead of to earth. When the first pulse arrives, after the application of h.t. to the circuit, V_2 only will be triggered because of the lower trigger voltage on V_1. However, once V_2 is conducting its cathode potential will be raised to approximately 40V and the trigger potential V_1 is therefore raised by a proportionate amount. The second pulse on the trigger is therefore sufficient to trigger V_1 and the circuit is tripped to the other state.

The considerations in the design of the anode circuit are again the same as previously and those for the trigger circuit are identical to those for the trigger of V_1 in Fig. 41.

It will be noted that the input pulse is applied to both the conducting and non-conducting tubes simultaneously. The conducting tube does not seriously attenuate the input pulse even when the source impedance is comparatively high. This is probably due to the fact that the input impedance of the conducting tube is momentarily high until the ionization level in the trigger cathode gap has been built up. Resistors in series with each trigger can be inserted to reduce any effects of the conducting tube on the input pulse. However, these resistors will effectively be in series with the signal source impedance and the cathode resistor and care must be taken to ensure that the discharge is not so limited that it cannot transfer to the anode.

In all three of the circuits described an output pulse for driving a succeeding circuit can be obtained by inserting a resistance R_k in series with one of the two cathodes. The form of the pulse is shown in Fig. 42 and the pulse height is given by the expression

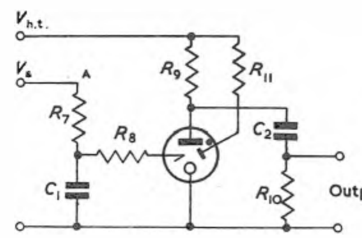

Fig. 43. Constant frequency oscillator

$$V_{out} = (V_{ht} - V_m) \cdot \frac{R_k(R_3 + R_4)}{R_3R_4 + R_kR_3 + R_kR_4}$$

V_{out} in general, should not exceed $\frac{1}{2}(V_{ht} - V_m)$ or the mechanism of extinction may be affected. Secondly, R_k limits the trigger discharge and for this reason should not exceed 6kΩ. However, even with these restrictions an output pulse in excess of 60V is readily obtained.

Self Extinguishing Circuits

In this section it is proposed to discuss two particular applications using the self extinguishing circuits; a constant low frequency oscillator and a thyratron firing circuit. The two circuits have been selected since they both require a tube with a stable firing characteristic, but it should be stressed that these are by no means the only applications for this type of circuit.

The Relaxation Oscillator

The complete circuit is shown in Fig. 43 and is basically the self extinguishing circuit described earlier, in which both the trigger and the anode are made self extinguishing. A short duration negative pulse output at a constant frequency f is obtained across the resistor R_{10}. A similar positive pulse could be obtained across a resistor connected in series with the cathode. The frequency range of the circuit is from a few pulses per minute to about 300c/s and the stability will be similar to that quoted for the timer circuits; (i.e. long term stability of 2 per cent).

The frequency of oscillation is given by the expression

$$f = 1/R_7C_1 \log \frac{V_s - V_o}{V_{t(ign)} - V_o}$$

where V_s is the potential at point A

and V_o is the potential at the trigger immediately after an anode cathode discharge (approximately 90V for the Z803U).

It will be seen that the above expression is the reciprocal of that for the timing period in an *RC* timer and for this reason a number of the considerations in the design of the trigger circuit are identical to those already discussed in the timer section: namely the factors determining the choice of R_T and C_1, the minimum frequency of operation (i.e. the maximum timing period), the stability and finally the method for compensating for tube to tube spread in $V_{t(ign)}$ and V_o.

The factors determining the maximum frequency are, however, different from those effecting the minimum timing interval. At the higher repetition frequencies the level of ionization in the tube just prior to the instant of break-

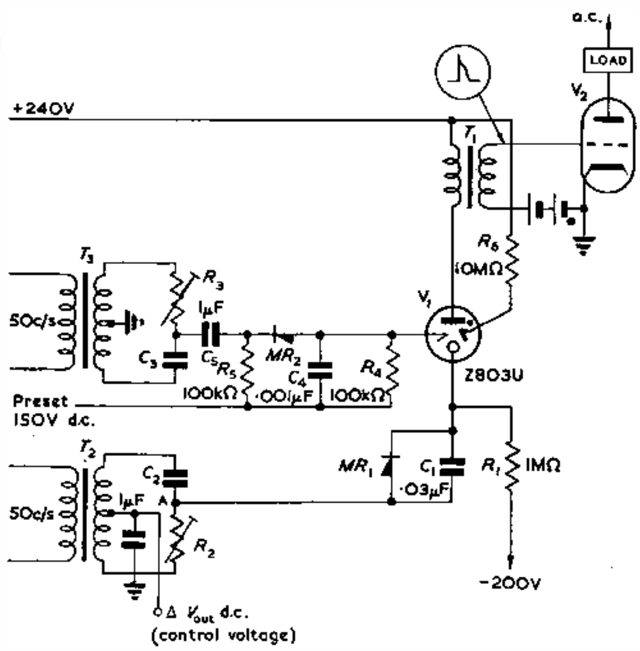

Fig. 44. Basic control circuit

down will be greater than that provided by the priming discharge. This is because the tube will not have completely deionized, although it will have recovered complete control. For that reason variations in the ionization time will be considerably less than normally experienced and the errors due to this cause can be neglected.

The maximum frequency limit is in fact set by the deionization time of the tube. It will be found from the curve of Fig. 19, published earlier, that the minimum time-constant for the anode circuit R_9C_2 is 1·7msec. It sets the upper frequency to a maximum of 300c/s., since to ensure that the anode potential is returned to substantially h.t. before each breakdown the trigger time-constant should be made greater than that of the anode.

Although the frequency of the output pulse is determined entirely by the trigger circuit the energy is directly proportional to the anode capacitor C_2. For values of C_2 greater than ·005μF the output pulse height is substantially independent of C_2 and only the duration of the pulse will increase with C_2.

As already mentioned in the section on the self extinguishing circuits the resistor R_{10} should not be greater than 1kΩ, and also if C_1 is less than ·0039μF the resistor R_4 should be omitted.

Firing Circuit for Grid Controlled Rectifiers

Thyratrons, excitrons, etc. are used in large numbers in a variety of power control applications. The method of controlling the power in the load is to vary the angle for which the rectifier conducts. The advantage of this system is that a large change in output power can be obtained with a small control power.

The most satisfactory method of control is by applying short duration pulses of variable phase-angle to the rectifier control electrodes or grids. However, in most industrial control systems the basic input signal is a variable d.c. voltage and the satisfactory conversion of such a signal into a variable phase-angle pulse has proved difficult.

The trigger tube, with its sharp transition from 'off' to 'on' is well suited to this type of duty and the use of a stable close tolerance tube leads to a circuit design in which

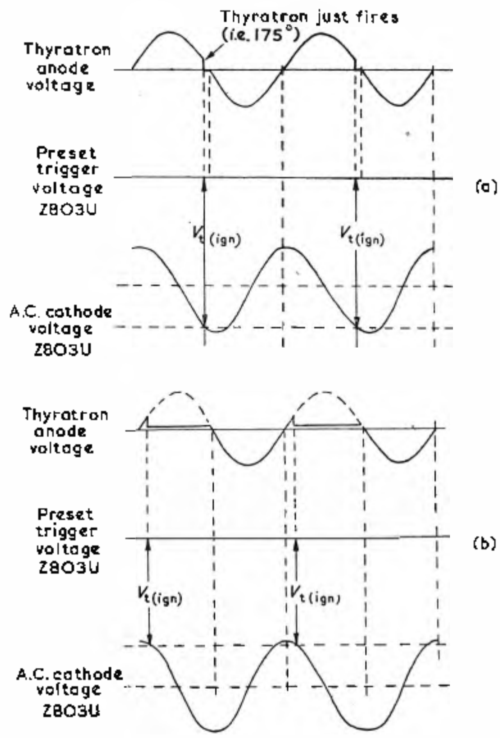

Fig. 45. Waveforms for circuit of Fig. 44

the input power needed to give a phase shift of 180° is very small, thus giving a high overall gain.

Fig. 44 shows a firing controller which is essentially a self-extinguishing circuit. When the tube is triggered the capacitor C_1 is discharged through the tube and the primary of transformer T_1. The current flow through the primary of T_1 produces a sharp voltage pulse across the secondary which is employed to trigger the thyratron V_2.

These pulses are synchronized to the mains by an a.c. voltage superimposed on the controlling d.c. voltage at point A. It is assumed for the present that the trigger is held at a constant potential. Fig. 45 shows the waveforms for the circuit at the two extreme settings of the d.c. control voltage showing that a complete 180° phase shift can be obtained.

It will be seen that in the case when the thyratron is fired at 0° the controlling voltage during the remaining whole cycle will be more negative than at the instant at which the tube is fired. Therefore, once the capacitor C_1 has discharged via R_1 to the potential at A the tube will fire again, earlier than required. To avoid this and thus to permit the full 180° phase shift a negative half sine wave is applied to the trigger via the rectifier MR_2 phased as shown in Fig. 46. This reduces the trigger cathode potential below the breakdown voltage during this period, but has no effect on the performance of the circuit when the tube is required to fire.

The rectifier MR_1 performs a number of functions. The primary duty is to chop the exponential cathode waveform and thus obtain a triangular waveform. Unless this is done a shorter time-constant must be used to ensure that the cathode potential returns to the potential at A within one cycle. However, with the short time-constant the circuit is more likely to give the second firing described earlier.

Secondly, the rectifier provides a low impedance path for the priming discharge which would otherwise produce errors due to the voltage drop across the cathode resistor.

It will be seen that the values of components in the

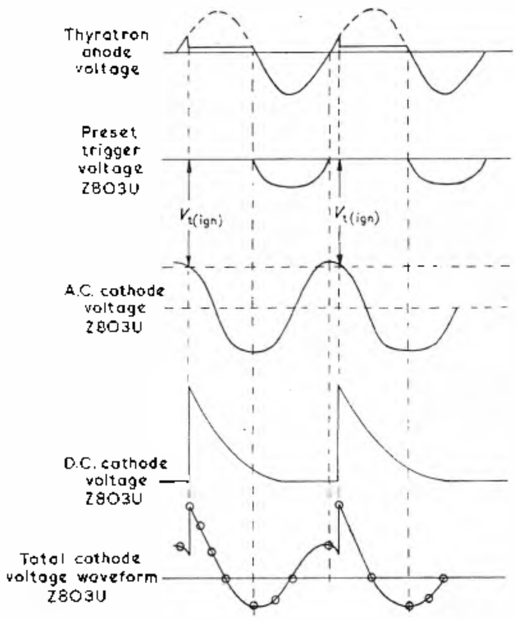

Fig. 46. Waveforms for circuit of Fig. 46 with negative half sine wave applied to trigger

trigger circuit are not such as to make the trigger self-extinguishing but the rise in cathode potential ensures the trigger is extinguished.

Capacitor C_2 and resistor R_2 determine the phase of the cathode voltage while R_3 and C_3 set the phase of the trigger voltage.

Power Supplies

In conclusion it will be useful to describe the design procedure for the anode supplies of circuits such as the two timers described earlier, since a considerable percentage of trigger tubes will be used in apparatus which do not require other supplies. A power supply for this type of instrument must be able to cope with the wide spread of nominal mains voltages met in this country, i.e., 200 to 250V.

A basic circuit is shown in Fig. 47, it consists of a half wave rectifier connected to the mains and a divider chain R_1 and R_2. The ratio of R_1 and R_2 are selected so that when the mains is at its highest value and the tube is non-conducting the output voltage does not exceed the maximum anode voltage of the tube. Secondly when the tube is conducting the supply must be capable of providing the required current. For calculations on the latter condition the effective source resistance of the rectifier must be known.

A very close approximation to the source resistance can be obtained by drawing a tangent AB (Fig. 48) to the output voltage-current characteristic of the rectifier and capacitor to be employed. They can then be regarded as a d.c. voltage source of E volts (approximately equal to V_{peak} of the supply) and a series resistance R_s equal to the slope of the line AB.

R_1 and R_3 can then be determined from the following two equations.

$$\frac{E_1}{R_1 + R_2 + R_s} = \frac{V_{a(max)}}{R_2}$$

$$\left[R_L + \left(\frac{R_3(R_1+R_s)}{R_1+R_2+R_s}\right)\right] i_m = (E_2 - V_m)$$

where $V_{a(max)}$ is the maximum anode voltage for the tube.

E_1 is the value of E at the maximum mains voltage.

i_m is the minimum necessary tube current for the application.

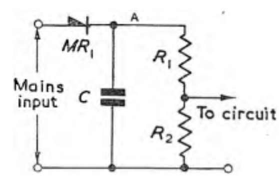

Fig. 47. Basic power supply

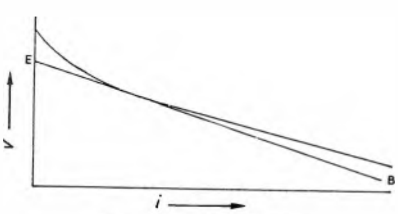

Fig. 48. Rectifier curves

V_m is the maintaining voltage of the tube.

E_2 is the value of E for the minimum mains voltage.

R_L is the anode load.

In cases where the anode load is purely resistive R_L can be zero and R_1 and R_3 made up to the required effective resistance. In this way the power dissipation in the divider chain is reduced.

A high voltage is available at point A for supplying other circuits such as the stabilizers in the timer circuit of Fig. 24.

When the anode load consists of a self locking relay the alternative circuit shown in Fig. 49 is more economical. The circuit consists of a half wave rectifier with its efficiency reduced by a series resistor R_3.

When the tube is triggered the capacitor C_1 momentarily supplies the current which is needed to pull in the relay and to close contact A_1. The supply now has only to provide the hold-in current of the relay and the voltage at point B can fall to the relay operating voltage.

The design procedure is slightly different from the previous case. The output voltage-current curves for a number of values of the series resistance are shown in Fig. 50. The curves have been drawn for a mains input voltage of 220V. For input voltages above and below this value the slope remains approximately the same but the curve is displaced so that it crosses the zero current axis at the appropriate peak voltage.

A curve is first selected which, at the minimum mains

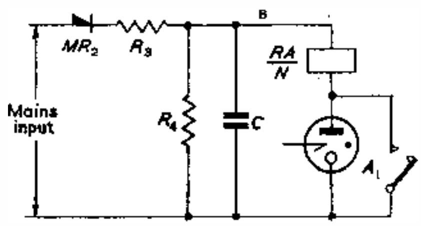

Fig. 49. Alternative power supply

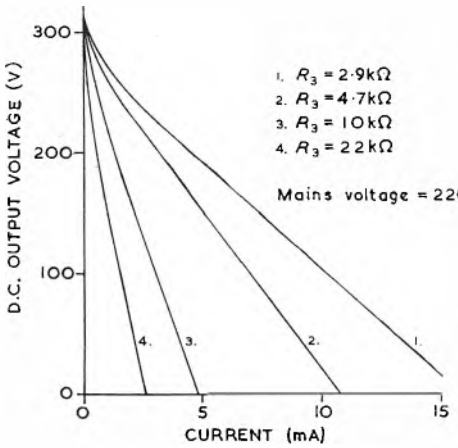

Fig. 50. Output voltage curves for Fig. 49

input voltage, gives the hold-in current i_R of the relay. Under this condition the voltage at B is equal to the normal relay voltage drop $i_R\, R_A$ (where R_A is the relay resistance). The curve for the selected value of R_3 at the maximum mains voltage is now drawn. The current i_B which must flow through R_4 to give an output voltage less than the maximum tube anode voltage, can then be obtained. R_4 is calculated such that this current flows at the maximum tube anode voltage

$$\text{i.e., } R_4 = V_{a(max)}/i_B$$

A check should be made to ensure that the output voltage exceeds the minimum anode voltage when the mains voltage is at a minimum.

It should be noted that the curves have been drawn for an ideal rectifier. When a metal rectifier, with a finite back resistance is employed the curves at zero current will pass through a voltage less than the peak supply volts.

Acknowledgments

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A New Process Control Installation

The recently installed control installation for continuous diffusers at the British Sugar Corporation's factory at Wissington in Norfolk, is believed to be the first closed loop electronic process control installation applied to the process of extracting sugar from sugar beet, in the world.

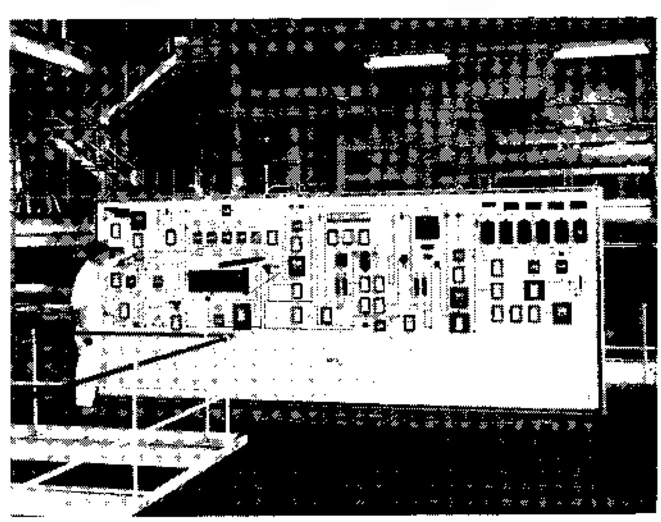
It was designed by Evershed & Vignoles Ltd, in conjunction with the British Sugar Corporation.

The installation makes use of a variety of transducers for flow, level, temperature, etc., which provide an electric signal which is either applied to Evershed 3-term process controllers, or used as an input to Evershed Simple Analogue Computers. Several of these computers, which are completely integrated into the installation, are employed to provide an output signal which represents the continuous evaluation of the equation governing the part of the refining process to which they are applied. These computers, in conjunction with the three-term mode of process control, permit extremely fine control of the manufacturing process which will further improve the performance of an already outstanding and up-to-date plant.

The graphic control panel which forms the heart of the whole installation, embodies a number of Evershed Miniature Recorders which will provide a permanent record of the performance of the plant, and will also furnish data for accountancy purposes etc. The control panel also includes Miniplaques, so that bumpless transfer from automatic to manual operation of the plant, when desired, can easily be accomplished. An automatic alarm system, covering every vital part of the plant, also terminates at the panel.

In view of the complexity of the process controlled, use has been made of the Evershed "In-Line" Scanner, which facilitates the supervision of the whole of the plant by a single operator, and which pin-points the location of any possible irregularity at once.

An interesting feature is the facility with which future extensions of the plant may be added to the present installation and incorporated at the control panel, without interference with the existing control arrangements.



The control panel at the Wissington factory

AN AMPLIFIER FOR A.C. BRIDGES

By A. H. Allan*, J. R. Gabriel* and B. H. Robinson*

The requirements of an amplifier for a.c. bridges working to an accuracy of one part in ten thousand or better, from power frequencies to audio frequencies, are discussed. An amplifier to meet, in part, these requirements, is described, and figures are given for its performance.

BRIDGES to be used to one part in ten thousand or better need precise balancing of the bridge. The attenuation through the bridge when it is just perceptibly off balance is well over 100dB. Since the voltage that can be applied to the unknown is limited, and the sensitivity of the detector is also limited, the amplifier must have a considerable gain, and the output due to thermal noise, to mains hum, and to stray voltages picked up by the circuit must be kept very low.

The obvious way to reduce noise is to filter out all frequencies except the one at which the bridge is working. To do this, a filter of exceedingly narrow bandwidth, variable from below 50c/s to the upper limit of audibility would be wanted. Some sensitive detector-amplifiers saturate at low input levels, so that when a bridge is more than a short distance from balance, it is impossible to tell whether adjustments are being made in the right or the wrong direction. Ideally, the detector-amplifier should show a continuous increase of output as the bridge is varied from balance to complete unbalance. That is to say, it should never saturate. The gain control which is usually fitted to prevent working in a saturated condition, is a nuisance which is minor but significant. Bridges of this accuracy are commonly built up from components as they are required, and may have output impedances of widely different values. The input transformer, which matches the bridge to the amplifier, should be easy to change, but should have connexions that do not readily pick up stray fields. It is easier to design a number of transformers with different ratios and impedances than a single one with a number of primaries.

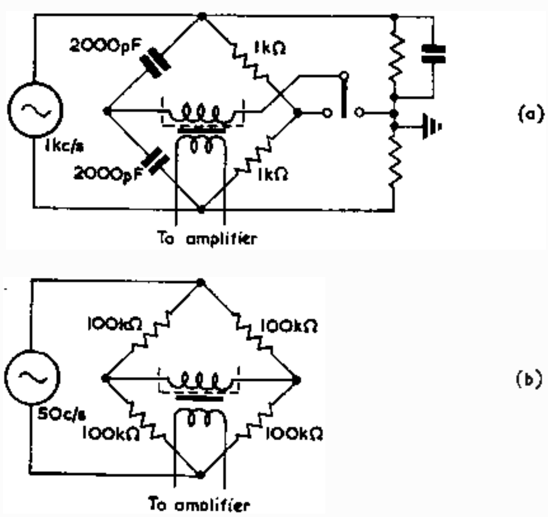


Fig. 1 (a) Schering bridge. (b) Resistance bridge

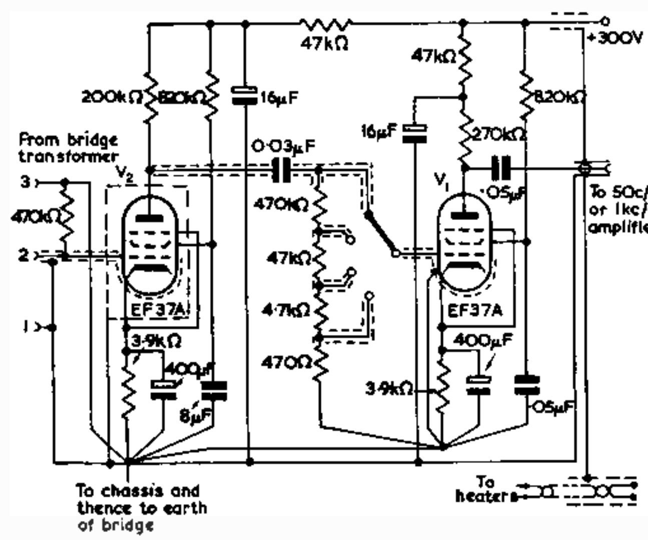


Fig. 2. A.C. bridge pre-amplifier

Description of Amplifier

The amplifier which is described in this note was intended only as a step towards the ideal. It was to be used with bridges, of which the two shown in Fig. 1 are typical; although it might be wanted for any frequency up to the top of the audio range, 50c/s and 1kc/s were the two most important frequencies. Saturation is prevented by a switched attenuator with steps of 20, 40 and 60dB. This arrangement does not introduce the noise of a continuously variable potentiometer, and small steps are not needed since the whole system from amplifier input to the detector (the ear, a vibration galvanometer or an oscilloscope) can tolerate a 20dB variation without overloading.

To attain flexibility, there are three units; a main amplifier, with a frequency range from 30c/s to 20kc/s, a frequency selecting unit with some gain, and a power supply. Two different frequency selectors have been made. One has a range from 40c/s to 60c/s and the other a narrow band centred on 1kc/s. The second can be switched to work untuned, and then has a range from 20c/s to 5kc/s.

The main amplifier (Fig. 2) is designed to have low noise. It is conventional except that the first valve is screened with 18 gauge copper, and the input lead is enclosed in copper tube. The heater leads can be connected either to the 6·3V supply of the power pack or to a battery. When the mains supply is used, the hum is reduced by an earthed variable potentiometer in the filament supply.

The best arrangement for earthing is one that uses no common earth lead for points of the circuit that work at very low levels and at higher ones. The actual earth used should be independent of the mains supply earth. The three chassis of the amplifier were connected together and then to the bridge earth by the screen of the input lead. The earth currents in this screen apparently cause no trouble, this lead is a screened twin, and the screen does not carry the signal current. Within the pre-amplifier the earthing arrangements are as shown in Fig. 2. The various earth connexions of each of the two stages are brought to a common point, and the two points are joined by a single wire.

The 1kc/s and the 40 to 60c/s amplifiers are designed to

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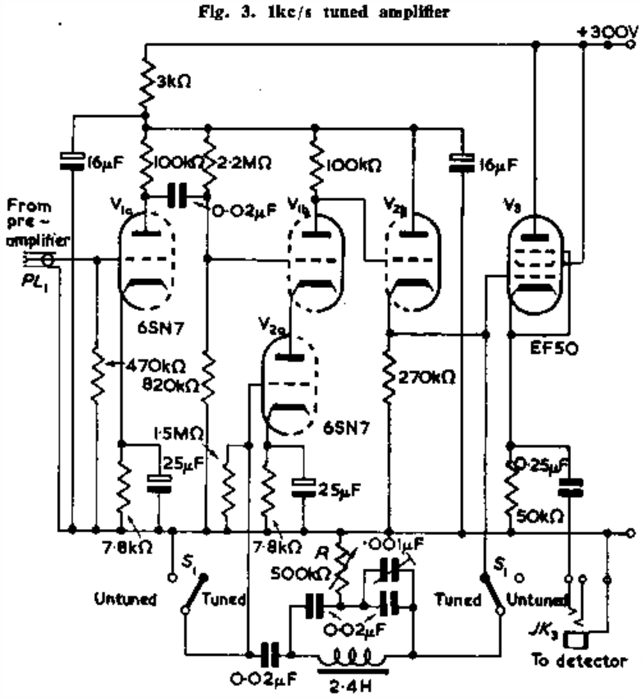
cover the two frequencies that are far the most commonly used. The bandwidth at 1kc/s is about 2c/s and at 50c/s about 0·6c/s. As a consequence, the noise present in the output is proportionately greater at about 50c/s than at 1kc/s. A practical limit to the useful gain of an amplifier is that it should give some definite noise output at full gain. In this case, the definite output that is wanted is fixed by the highest level that an observer can conveniently tolerate in headphones, and an observer who is constantly listening for exceedingly faint sounds at balance cannot tolerate very high maxima. The gains of the two frequency selective units have been designed to give the highest tolerable noise output when working at full gain. Since the low frequency amplifier is inherently much the noisier of the two, the gain that it can have has had to be kept much lower than that of the 1kc/s one.

The 1kc/s amplifier (Fig. 3) has a bridged-T network in a negative feedback line, and follows Valley and Wallmann[1]. Since its input level is relatively high, no special precautions need be taken about screening or earthing. It has two adjustments:—a feedback control which is to be set so that the amplifier is a little short of oscillation, and a fine tuning control, adjustable over about 10c/s. Normal tuning is of the supply oscillator to the amplifier frequency.

The 40 to 60c/s amplifier (Fig. 4) uses a twin-T network, and the tuning frequency is variable over a comparatively wide band. The three feedback resistors R_1, R_2, and R_3 must be separately adjusted to the working frequency by connecting an oscilloscope in the feedback path at JK_2, and varying the three resistors until the twin-T network attenuates most at the working frequency.

The power supply is conventional and is not illustrated.

Fig. 3. 1kc/s tuned amplifier



Performance

The minimum voltage that the wave analyser used could measure was 5mV. Zero means a voltage of less than this value. Voltages expressed in dB are referred to 1V as 0dB.

Pre-amplifier

Input resistance 470kΩ
Output load resistance 470kΩ

Noise Output

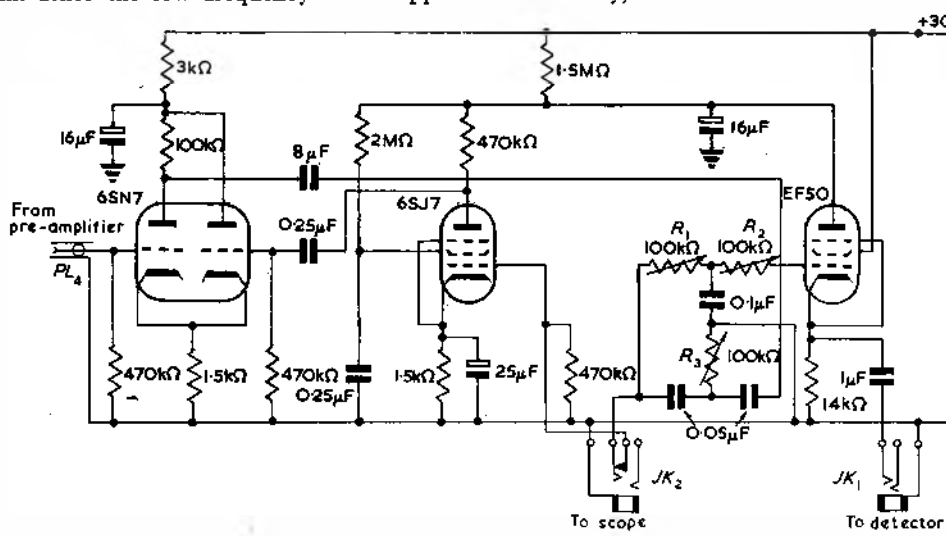


Fig. 4. 40 to 60c/s tuned amplifier

	Gain setting 0 (max) (mV)	Gain setting 60 (min) (mV)
(1) Random Noise (heaters supplied from battery)		
Input open	135	0
Input shorted	40	0
Theoretical, due to input resistance	100	0·1
(2) Mains Noise (i.e. 50c/s component of output voltage		
Heaters and h.t. supplied from mains		
Input open	830	0
Input shorted	75	0
Heaters supplied from battery; h.t. from mains		
Input open	0	0
Input shorted	0	0

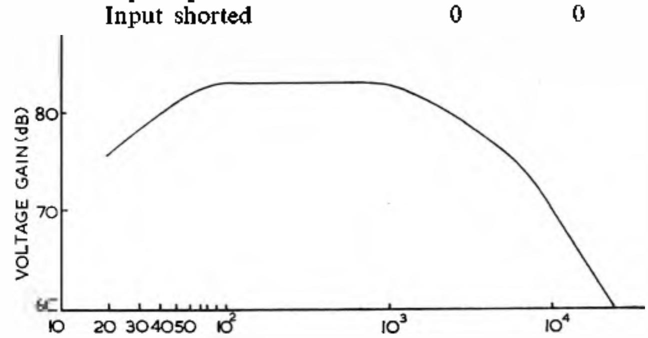


Fig. 5. Pre-amplifier characteristic

When the heaters were fed from mains the random noise voltage remained unaltered, and was superimposed on the mains-frequency output.

Gain-frequency Characteristic (Fig. 5)

Gain control set at 0 (max) and output brought to a constant level of 1V (0dB).

The gain-frequency characteristic is not affected by the gain control setting provided the amplifier is not overloaded.

GAIN CONTROL SETTING	ATTENU-ATIONS	MID-BAND GAIN (DB) (300c/s)	APPROX. OUTPUT AT START OF SATURATION (V)	APPROX. INPUT (mV)
0	0	83	34	3·2
20	20	63	56	55
40	39	44	58	560
60	61	22	4	340

1kc/s Amplifier

Input resistance 470kΩ

Output resistance 1·2kΩ

Untuned Characteristics

Noise output: 20mV of main hum, with the input open or shorted.

Gain-frequency Characteristic (see Fig. 6)

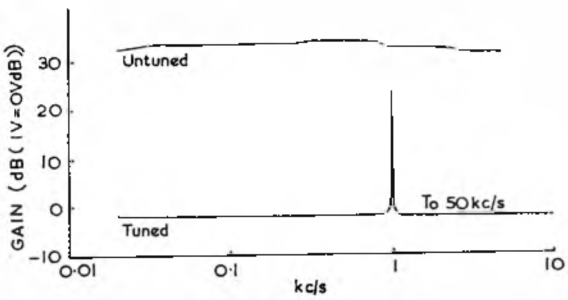


Fig. 6. 1kc/s amplifier characteristic

General

Approx. output at start of saturation 20V

Approx. input at start of saturation 430mV

Tuned Characteristics

Noise output: Noise output is less than 5mV with input open or shorted.

40 to 60c/s Amplifier

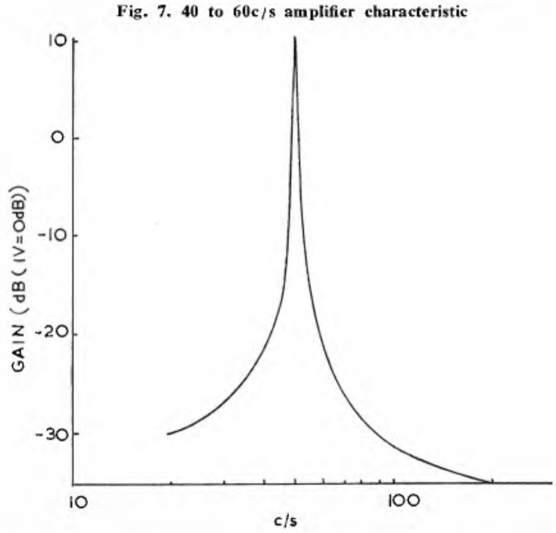
Input resistance: 470kΩ

Output resistance: 2·5kΩ

Characteristic: when tuned to 50c/s less than 5mV noise output under all conditions.

Gain-frequency Characteristic (see Fig. 7).

Fig. 7. 40 to 60c/s amplifier characteristic



Pre-amplifier with 1kc/s Amplifier

Noise Output

(1) Heaters and h.t. supplied from mains:

GAIN CONTROL SETTING	NOISE VOLTS AT OUTPUT			
	INPUT OPEN		INPUT SHORTED	
	MAINS	RANDOM	MAINS	RANDOM
UNTUNED:		(1)		
0	11V	4V	2·5V	0·65V
20	1·1V	0·4V	250mV	65mV
40	110mV	40mV	60mV	15mV
60	70mV	10mV	60mV	15mV
TUNED:		(2)		
0	200mV	90mV	50mV	10mV
20	20mV	9mV	5mV	0
40	5mV	0	0	0
60	0	0	0	0

(2) Heaters supplied from battery and h.t. from mains:

UNTUNED:				
0	0	5V	0	1·3V
20	0	0·5V	0	130mV
40	0	50mV	0	13mV
60	0	10mV	0	5mV
TUNED:				
0	0	110mV	0	20mV
20	0	13mV	0	5mV
40	0	0	0	0
60	0	0	0	0

(1) Approx. output due to thermal noise from input resistance 2½V.

(2) Approx. output due to thermal noise from input resistance 4mV

Pre-amplifier with 40 to 60c/s Amplifier

(1) Heaters and h.t. supplied from mains:

Noise output: when tuned to 50c/s, less than 5mV of random noise at any gain setting with input open or shorted.

GAIN CONTROL SETTING	MAINS NOISE VOLTS AT OUTPUT	
	INPUT OPEN	INPUT SHORTED
0	0·4V	110mV
20	40mV	15mV
40	20mV } Approx.	0
60	5mV } Approx.	0

With input open and gain set at 0 (max)

AMP TUNED TO C/S	MAINS-NOISE OUTPUT
40	15mV
50	0·4V
60	15mV

(2) Heaters supplied from battery and h.t. from mains:

There was no mains or random noise output under any conditions of input, gain setting, or amplifier tuning.

Notes:

(1) Voltages are expressed as $1/\sqrt{2}$ of peak values.

(2) When the heaters were fed from mains, the earthing potentiometers on the power supply were always set to give a minimum of mains-frequency output.

(3) When the heaters were fed from a battery, one pole was always earthed.

Acknowledgments

Acknowledgment is due to Mr. W. H. Ward, under whose direction this work was carried out, for his helpful suggestions.

REFERENCE

(1) Valley and Wallman. Vacuum Tube Amplifiers. (McGraw Hill).

A U.H.F. Wide Band Amplifier

By J. Kason, A.M.I.E.E., A.M.Brit.I.R.E.

A wide band u.h.f. amplifier of a mid-band frequency of 500Mc/s and having a bandwidth equal to approximately 40Mc/s is analysed. The anode and grid impedances of a stage are evaluated and a method of coupling, using a balanced delay line in the anode, is shown. The gain and noise factor are also considered.

THE design of a wide band u.h.f. amplifier is analysed, and a lay-out diagram, and its electronic equivalent are shown.

Wide band amplifiers for wire distribution in Bands 1, 2 and 3 and their design have already been described in the past[2,6,7]. With the continuous growth of programmes in Band 3, a need may arise in the near future for the distribution of new television black and white or colour channels on higher frequencies. The amplifier described below could meet such a demand.

Circuit Arrangement

A single stage amplifier was investigated and later two stages with staggered circuits and with circuits tuned to the same frequency (synchronous) were considered. The circuit diagram of a two stage amplifier is shown in Fig. 1 and Fig. 2 shows the main details of the assembly.

Examination of Fig. 1 shows that balanced delay lines are used in each anode circuit and to couple the two valves. Parts of the lines are shown dotted to indicate that one half exists only as an image in the metal of the case. In addition, certain other components which do not exist as separate physical entities are also shown dotted.

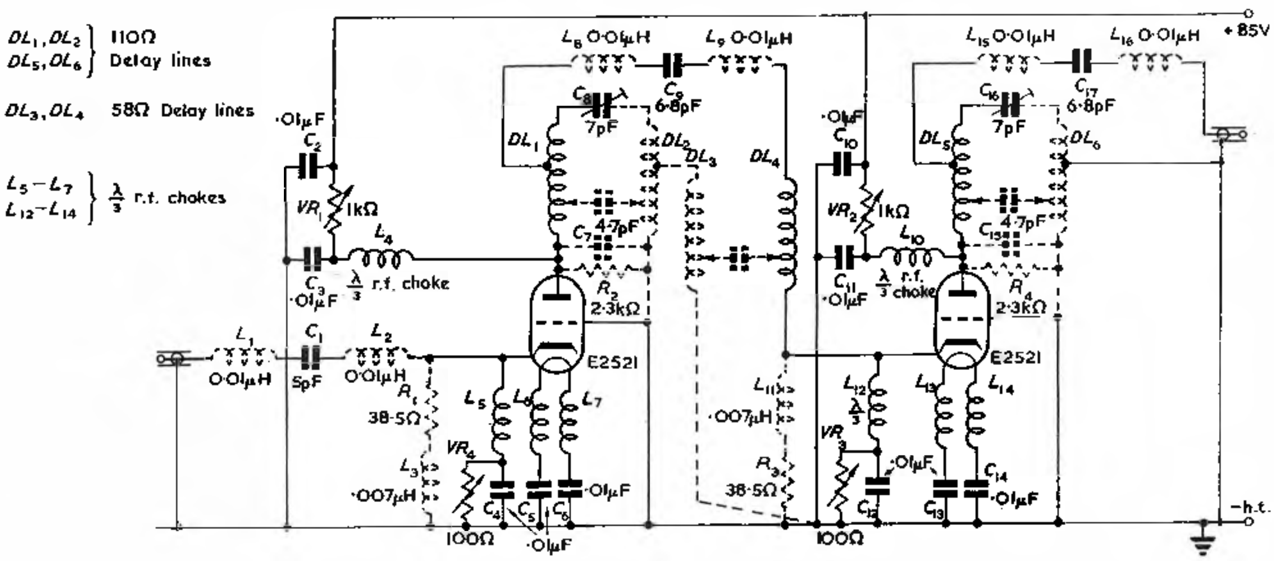


Fig. 1. Two stage wideband amplifier

Design Considerations

It is extremely difficult, at u.h.f., to keep the anode bandwidth below 10Mc/s, and as the figure of merit of a valve is approximately constant, the gain per stage is small. It was decided to utilize this wide bandwith in the design of the amplifier, but keeping the noise figure as small as possible. The optimal noise bandwith per stage was calculated, and the design of the amplifier was centred around that figure.

A number of different valves were considered with regard to signal-to-noise ratio, and the best compromise, consisting of two E2521 valves in the circuit shown in Fig. 1 was adopted. The compromise arises because, although planar valves 416A, 6299, 5768, 6BY4, etc. have a better signal-to-noise figure, it was considered uneconomical to adopt their costly mechanical construction for commercial usage.

The E2521 is a miniature type valve, which requires a B9A base, has a planar construction of electrodes for operation up to 1 000Mc/s. The valve must be operated with −0·5V on the grid and 15mA anode current exactly, in order to obtain optimum signal-to-noise ratio.

Anode Circuit

The anode circuit consists of a 110Ω balanced delay line, the physical portions of which are a copper rod and a trimmer capacitor as shown in Fig. 2. A calculated electrical zero appears at approximately 2·19in from the trimmer and a 58Ω tapping point was calculated to be 0·72in from the zero point. The trimmer is at the far end of the line and offers a convenient tuning facility. The delay line tunes with the anode stray capacitances which are equal to 4·7pF, viz C_{ak} 1·8pF + C_{stray} 0·2 + C_{choke} 2·7.

Anode damping is provided by an impedance r_a', shunted by the transferred load impedance, and the valve operates in the maximum output power condition. r_a' is assumed to be $\frac{1/g_m + 1/R_L}{1/g_m} r_a$, where r_a is the anode impedance, and is further modified to r_a'' by the capacitances at each end of the line. Under the conditions shown, r_a'' becomes 21·2kΩ. The calculated value of the transferred load impedance is 2·62kΩ so that the total effective dynamic impedance of the anode tuned circuit R_a, is 2·33kΩ and this results in a bandwidth of 28Mc/s.

The 6·8pF coupling capacitors (C_9,C_{17} in Fig. 1) form series resonant circuits with their own leads to provide minimum phase shift. The anode line phase shift is approximately 1 radian (see Appendix).

The amplifier could be further improved by connecting the h.t. choke to the electrical zero point instead of to the anode but, because of the difficulty of knowing the exact position of the zero point, this has not been done at the present stage.

INPUT CIRCUIT

The input impedance Z_{in} consists of the input resistance R_{in} shunted by the reflected valve and load impedances and

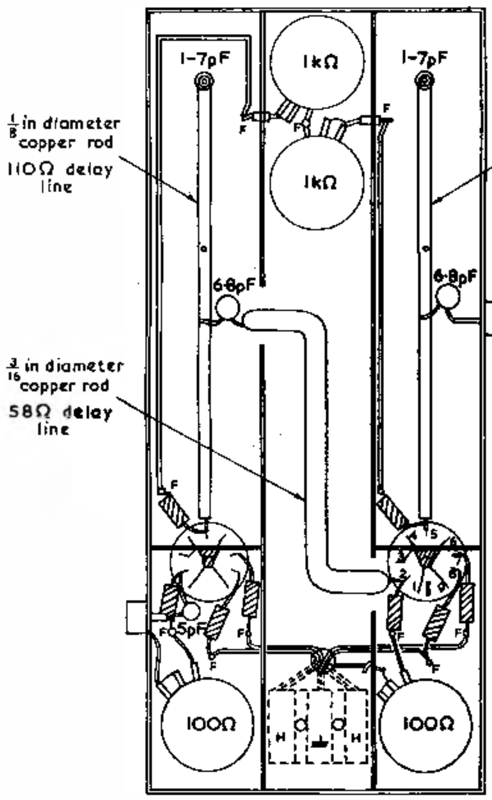


Fig. 2. Main details of assembly

by C_{gk} all in series with the 5pF capacitor (C_1 in Fig. 1) and with the inductance of its leads. The resultant calculated input impedance (see Appendix 2) at the socket is approximately 60Ω.

GAIN[5,7]

The gain of a single stage, two valve, amplifier was calculated with the following results:

(1) A staggered pair with each valve circuit having a bandwidth of approximately 36Mc/s, a bandwidth factor of 0·06 and a displacement factor of 1·03, was found to have a gain of 18dB for a total bandwidth of 40Mc/s at the 3dB points (See Appendix 3).

(2) A synchronous cascaded, two stage, two valve amplifier having a bandwidth per stage of over 50Mc/s is required for an overall bandwidth of 38Mc/s at the 3dB points and this has a total gain of 16dB.

NOISE (See Appendix 4)

The noise factor for an E2521 valve is 8·9dB at 500Mc/s. The mid-band (500Mc/s) noise factor is further increased by the addition of a second stage, since:

$$F_\varepsilon = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_2 + \ldots\ldots$$

where $F_1 F_2$ = Noise factors

$G_1 G_2$ = Mid-band gain of individual stages.

Hence, for a staggered pair, for the worst signal-to-noise point, viz. 520Mc/s:

$$F_\varepsilon{}^1 = 7{\cdot}75 + 7{\cdot}75/4 = 9{\cdot}69 \cong 9.85\text{dB}$$

and the equivalent noise at the input becomes:

$$e_n{}^2/R_{in} = F_\varepsilon{}^1\, kT(BW) + k\beta T(BW)$$

$$\therefore\ e_n = [kT(BW)\,(F_\varepsilon{}^1 + \beta)\,R_{in}]^{\frac{1}{2}} = 3{\cdot}78\mu\text{V}$$

Similarly, the noise figure for two cascaded stages was evaluated.

$$F_\varepsilon{}'' = 7{\cdot}75 + 7{\cdot}75/6{\cdot}3 = 8{\cdot}9 \cong 9{\cdot}5\text{dB}$$

and the corresponding noise figure at the input becomes 3·67μV.

The optimum signal-to-noise bandwidth per stage for an E2521 valve working at 500Mc/s with the circuit shown in Fig. 1 has been calculated to be approximately 33Mc/s.

General Description

The amplifier is suitable for working into a 60Ω load, which is tapped across the output delay line. The two valves are inter-connected through a 58Ω delay line via an isolating capacitor which, together with its leads, behaves as a series tuned circuit.

The amplifier, whether having staggered circuits or not, has the same components, the only difference being the frequencies to which they are tuned.

The staggered pair has its first valve anode circuit tuned to 485Mc/s (BW = 29·1Mc/s) and the second 515Mc/s (BW = 30·9Mc/s).

The cascaded stages are both tuned to 500Mc/s, the bandwidth for each being 55Mc/s and for the two 38Mc/s at the 3dB points.

Conclusions

The u.h.f. wide band amplifier, intended for wire distributing has been proved. Although valves such as the 6BY4 will provide an improvement of 3 to 4dB on the noise factor, they require concentric lines and the resultant complex mechanical construction adds to cost. The E2521 was chosen as the best commercial compromise.

The amplifier is simple to make, to align, and to maintain.

Acknowledgments

The author wishes to thank the M.O. Valve Co. for their helpful suggestions and supply of an E2521 valve.

APPENDIX

(1) ANODE CIRCUIT[1,6]

The tuning of the anode circuit is evaluated below:

$$Z_s = Z_o \frac{Z_R + Z_o \tanh Pl}{Z_R + Z_o \tanh Pl}$$

$$\simeq Z_o \frac{Z_R + jZ_o \tan\Theta}{Z_o + jZ_R \tan\Theta}$$

$$\simeq Z_o \frac{Z_R + jZ_o\Theta}{Z_o + jZ_R\Theta} \qquad \text{for } \Theta < \tfrac{1}{2} \text{ radian}$$

$$\Theta = 2\pi\, l/\lambda, \text{ where } l = \frac{\text{physical length}}{\text{velocity factor}}$$

and $2\pi\, l/\lambda \simeq 1$ radian.

Now the diameter of the line is ⅛in and the distance from the chassis is 5/16in, so that

$$Z_o \simeq 110\Omega$$

The anode stray capacitances were shown to be 4·7pF, i.e. approximately $-j68\Omega$ at 500Mc/s.

$$\text{Thus } Z_s = \frac{110 - j68 + j110 \times 1{\cdot}785}{110 + j(-j)\,49{\cdot}7 \times 1{\cdot}785} \simeq +j68$$

Hence the trimmer value becomes equal to the total anode strays, viz: the anode resonance will be obtained with the trimmer setting of 4·7pF approximately.

(2) Input Impedance[1,4]

The impedance at the input terminals is evaluated as follows.

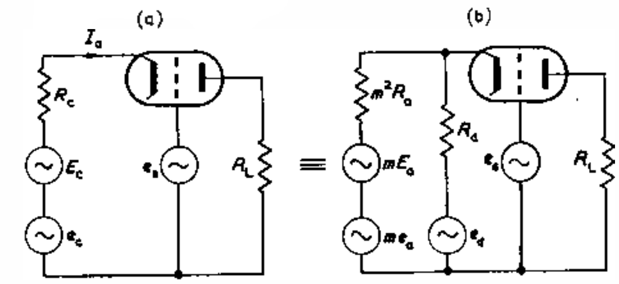


Fig. 3. Evaluation of input impedance

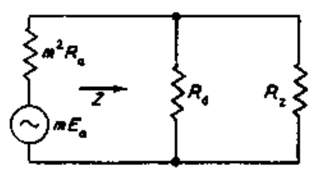


Fig. 4. Input for signal without noise

Sources marked with a large E are signal sources and those marked with a small e are sources of noise.

$$R_c = \frac{m^2 R_a R_d}{m^2 R_a + R_d} \text{ and } E_c = mE_a \frac{R_d}{m^2 R_a + R_d}$$

$$\text{Also, } I_a = \frac{(\mu + 1)\,E_c}{R_o + R_L} \text{ where } R_o = R_a + (\mu + 1)\,R_c$$

Hence, I_a =

$$\frac{E_c}{R_o + R_z}, \text{ where } R_z = \frac{R_a + R_L}{\mu + 1} \simeq 1/g_m + R_L/\mu$$

Thus, for a signal without noise, circuit Fig. 3 (b) can be shown to be as Fig. 4.

$$\text{and } R_z \simeq Z_{in} \simeq \frac{21200 + 2620}{41} \simeq 58\Omega$$

Slotted line impedance measurements made with simulated anode impedance gave $Z_l = 60 \tan 27°\Omega$.

(3) The Design of a Staggered Pair[5,7]

The design of a staggered pair having a bandwidth of 40Mc/s at the 3dB points and a gain of about 18dB at 500Mc/s, is briefly explained.

$$\delta = 1/Q = 40/500 = 0{\cdot}08$$

hence, the bandwidth ratio $d = 0{\cdot}06$ and the displacement $\alpha = 1{\cdot}03$, so that $f_1 = 500 \times 1{\cdot}03 = 515$Mc/s and $BW_1 = 515 \times 0{\cdot}06 = 30{\cdot}9$Mc/s. Also $f_2 = 500/1{\cdot}03 = 485$Mc/s and $BW_2 = 485 \times 0{\cdot}06 = 29{\cdot}1$Mc/s.

The staggered pair gives a Tchebycheff response curve. The improvement in $(G \times BW)$ over 2 stages, tuned to the same frequency, is

$$\frac{1{\cdot}2 \times n}{1{\cdot}1 \times n/2} \simeq 1{\cdot}54.$$

The gain per valve =

$$\frac{\text{Figure of Merit}}{BW} \simeq \frac{3{\cdot}1 \times 40}{30{\cdot}9} \simeq 12{\cdot}8\text{dB}.$$

$$Q_1 \text{ (anode)} = Q_2 \text{ (anode)} = F_1/BW_1 = F_2/BW_2 \simeq 16{\cdot}65.$$

(4) Noise[1,3,4,5]

The shot noise power at the output $= FkT(BW)G$,

where F = mid-band noise

G = mid-band gain

k = Boltzmann's constant

T = temp. in °k

BW = bandwidth

Also $F = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_2 + \ldots$

Hence the noise power at the input $\simeq FkT(BW)$. Therefore the total noise at the input becomes:—

$$e_n^2/R_{in} = FkT\,(BW) + k\beta T(BW)$$

where β = noise temperature ratio

e_n = noise voltage

R_{in} = input impedance

$$\therefore e_n = [kT(BW)\,(F + \beta)\,R_{in}]^{\frac{1}{2}}$$

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Closed-Circuit Television Aids Radiation Treatment

Closed-circuit television is being used in the Royal Marsden Hospital, London, to assist in the deep therapy radiation treatment of patients. The equipment is used in conjunction with a Telecaesium Ray Treatment Unit and permits the treatment to be carried out by remote observation. In this way doctors and radiographers are safe-guarded against excess radiation which can produce harmful effects.

Observation of the patient is necessary during this type of treatment since a beam of gamma rays is accurately directed into the affected region. It is therefore important that this region be kept in a fixed position relative to the beam, since a centimetre's movement in any direction would render the treatment useless.

By keeping the patient under constant supervision during the irradiation treatment any movement is immediately observed. The treatment is then discontinued while the patient is repositioned. In this way the specialist knows that the equipment is being used at maximum efficiency and in the most economical manner.

The television component has been designed so that no operating skill is required. It is brought into use simply by the operation of a master switch and normally no further adjustment is necessary.

The camera employed is the Marconi 1in industrial vidicon camera Type BD835, and the installation operates on a 625 line system. The 14in industrial monitor is an integral part of the telecaesium unit control panel.

A Millivoltmeter for Ultra High Frequencies

By C. C. Eaglesfield*, M.A., A.M.I.E.E.

A method is described for making measurements at frequencies up to a few hundred megacycles per second. The technique consists of using a modulated signal generator and having a crystal diode probe to detect the modulation.

THE author was recently faced with making measurements on mainly resistive networks at frequencies up to a few hundred megacycles per second, the only instruments available being a signal generator and a conventional valve-voltmeter. It was found that there was a very severe restriction in the small overlap between the greatest voltage available and the smallest measurable; in addition the capacitive loading of the valve-voltmeter was excessive.

An alternative technique was then tried, of using the signal generator modulated and having a crystal diode to detect the modulation, which was then amplified by a sensitive amplifier. This technique is common in standing-wave measurements, for which purpose special tuned low-frequency amplifiers are available.

It was immediately clear that the method was in essence a solution, since the sensitivity was greatly increased and the loading reduced. However between the first appreciation of a possible indicator and the realization of an instrument which can be dignified by the name of voltmeter, some development was needed.

The detected voltage from a crystal diode due to a small modulated r.f. voltage is proportional to the percentage modulation and to the square of the r.f. voltage. This square-law behaviour is not well presented on a linear scale, but meters exist with pole-pieces so shaped as to give a well-graded decibel scale.

The practical problem splits into two parts, the crystal probe and the amplifier.

Only three components are necessary in the probe, the circuit of which is shown in Fig. 1: a feed capacitor, the crystal diode and a shunt resistor. (The r.f. path to the crystal is completed by the cable capacitance.)

The most important requirements in the diode are small size, low capacitance and high resistance.

The shunt resistor should be of small-size and low capacitance; its magnitude should be intermediate between the crystal and the input resistance of the amplifier.

The magnitude of the feed capacitor is a compromise: by making it rather small, a probe of low capacitance and conductance is obtained, but there is a loss of sensitivity and low-frequency response.

The photograph (Fig. 2) shows an experimental probe before and after encapsulation. The metal ring is for the earth connexion; the flexible clip-lead shown is adequate up to about 300Mc/s.

The amplifier must have very great sensitivity and a meter scale with a good decibel range. The author has found it convenient to use a Furzehill Type V200 valve-voltmeter. This has a very good decibel scale and successive ranges increase by 10:1, i.e. 20dB. With the square-law detector, each range covers 10dB. The sensitivity of the standard instrument is as great as its wide band permits but the author has found it possible to increase the sensitivity considerably, by restricting the band to frequencies normally used for the internal modulation of signal generators.

This great sensitivity is of value, because it gives a wide range of voltages over which the diode response is square-law: this simplifies the calibration very greatly.

The narrower the frequency band, the greater the sensitivity can be, which suggests the desirability of tuning; however, the author was disappointed at the small advantage got in practice by tuning. The explanation seems to be as follows: for a good amplifier, the residual indication arises from noise in the diode and this noise voltage, in the absence of 'flicker effect', is proportional to the

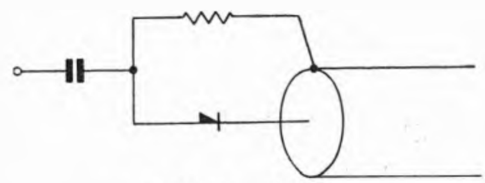

Fig. 1. The probe circuit

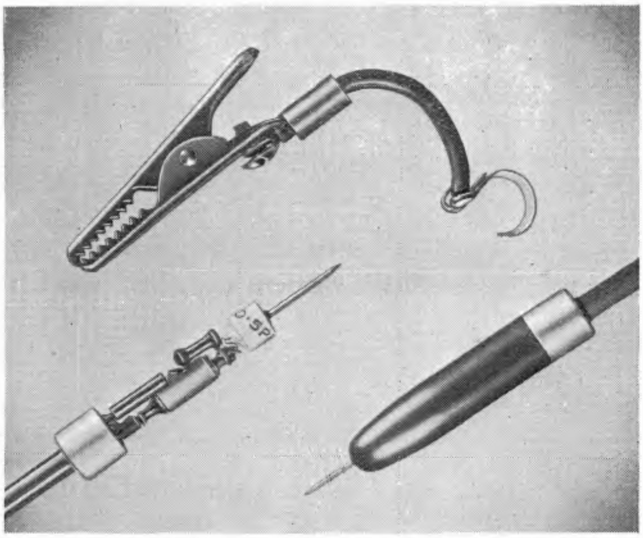

Fig. 2. An experimental probe, before and after encapsulation

square root of the bandwidth. The modulation, however, is proportional to the square of the r.f. voltage. It therefore follows that for a given 'signal-to-noise', the r.f. voltage is proportional to the fourth root of the bandwidth.

With a sufficiently narrow band, a fair improvement can be obtained, but the practical inconvenience is considerable. The author is of the opinion that it is hardly worthwhile to attempt to restrict the band more than has been mentioned above. Of course, if all signal generators had a standard, accurately set modulation frequency, it would be another matter.

As an indication of the possibilities of the method, some performance figures follow for a modified Furzehill voltmeter and experimental probes.

Input capacitance:	0·4pF
Input resistance:	0·25MΩ
Smallest voltage indicated:	3mV
Largest voltage indicated:	0·3V

* *Standard Telephones & Cables Ltd.*

The capacitance and resistance of the probe were measured at 50Mc/s. For the voltage measurements the modulation depth was 60 per cent; by largest is meant that square-law indication is preserved.

The performance at different frequencies is shown in Fig. 3. The standard used was a Marconi Ekco signal generator, Type 801B, and the voltage was measured at the end of the flexible cable, which was terminated in a 20dB pad. There is some undulation in the curve, indicating a small amount of standing wave in the signal generator cable, but on the whole the result speaks well both for the generator and the voltmeter. It will be noticed that the drop at the lowest frequency, 10Mc/s, is not large.

In conclusion, it can be said that while the method described has certain limitations and the sensitivity is less than can be got at low frequencies, yet it extends the frequency range over which millivoltmeters can be available by a couple of orders. It will be noticed how conveniently it takes over just where low-frequency millivoltmeters reach their limit.

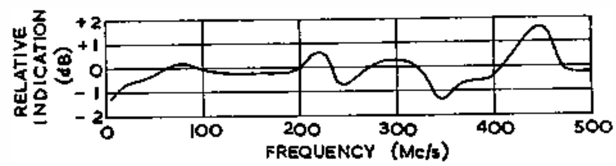

Fig. 3. The indication of the instrument as a function of frequency

Acknowledgments

Finally the author is pleased to acknowledge the help he has received from his colleagues and in particular the skilful construction by Mr. P. Barker of the probes; and to the Directors of Standard Telephones and Cables Ltd for permission to publish.

A Compact Stabilized D.C. Supply for Valve Heaters

By J. C. S. Richards*, B.Sc., Ph.D.

A stabilized supply giving up to 0·8A at 6·3V d.c. is described. Mains voltage variations of ±10 per cent affect the output by less than ±0·05 per cent, and internal resistance is less than 0·05Ω. The apparatus is designed to be incorporated in a stabilized h.t. supply, and the additional chassis space and rectified power required by it is comparatively small.

NUMEROUS circuits for stabilizing valve heater supplies have been described in the literature (see Benson[1]), but most of them are designed to feed a large number of valves and are consequently bulky. In many electronic instruments a few only of the valves require a stabilized heater supply. The circuit to be described gives up to 0·8A at 6·3V d.c.; this is sufficient for (say) four low noise pentodes, or two double triodes and a pentode, such as may be found in the first two stages of a high gain direct coupled amplifier. The apparatus is compact. If it is used in conjunction with a stabilized h.t. supply, as it is intended to be, it need occupy less than 20in² of chassis area.

Basic Principles

The complete circuit is shown in Fig. 1. The valve V_3 operates under class-C conditions as a reaction oscillator. Its frequency, nominally 20kc/s, is determined by the tuned circuit in its anode. The 20kc/s signal is rectified and smoothed by MR_1, L_1 and C_6 to give the 6·3V d.c. output. This output is compared with a fixed reference voltage and any difference between the two is amplified by the transistors X_1 and X_2 and the thermionic valve V_1, and applied to the grid of V_2. The valve V_2 acts as a cathode-follower and alters the mean screen grid and anode potentials of V_3 in such a way as to tend to keep the output voltage constant. The general theory of this type of feedback stabilizer is well known and need not be discussed here. Some details of the design, however, deserve comment.

The Transformer

Complete specifications for the transformer T_1 and the smoothing choke L_1 are given in the appendix. In general, the higher the frequency of oscillation, the smaller the physical size of transformer T_1 and the choke L_1. The upper limit to the frequency is set by the metal rectifiers, which become less efficient when the frequency is appreciably greater than 10kc/s. This frequency was initially chosen, but it was found difficult to prevent, completely, acoustic radiation from the components and chassis. A shrill whistle, although faint, can be very annoying, and the slight drop in rectifier efficiency when the frequency is raised to 20kc/s is more than compensated for by the resulting silence.

It is desirable to make the unloaded Q of the anode coil of T_1 as large as possible; by using a ferrite 'pot' core the eddy current losses and the external field are both made negligible. Since the alternating voltage across the coil at full output is approximately 150V r.m.s. hysteresis losses in the core are not negligible, and at a frequency of 20kc/s the 'proximity' effect must be taken into account when assessing the copper losses. The size of air-gap in the core and the number of turns and gauge of the winding have thus to be carefully chosen. The capacitor C_2 must have a mica or other low loss dielectric.

The Controlling Amplifier

The controlling amplifier (X_1, X_2, V_1 and V_2 in Fig. 1) is required to detect changes in its input voltage of less than 5mV. If thermionic valves (e.g. a balanced pair of triodes or pentodes) were used in the input stage; it would be necessary to feed their heaters from the stabilized 6·3V supply in order to reduce their drifts to less than 10mV. This is clearly not desirable in the present instance, when the total current available from the 6·3V supply is only 0·8A. Fortunately, the input voltage to the amplifier is developed across a comparatively low impedance, and a pair of junction transistors (X_1 and X_2) forms an ideal input stage. Drifts arising from temperature changes are minimized by selecting a pair of transistors with reasonably similar (within 20 per cent say) values for I_{co} and α', and connecting them in the circuit shown, in which the large value of common emitter resistance serves to stabilize the collector currents in a manner analagous to the action of a large common cathode resistor with a thermionic valve pair.

As an additional precaution, the transistors are set in holes drilled in an aluminium block which is mounted on grommets in the coolest part of the chassis. It is not difficult to site the block so that its temperature is not more

* *Natural Philosophy Department, Aberdeen University.*

than 10° C above room temperature. Since there are no high frequency currents associated with the transistor stage, there is no disadvantage in using long connecting leads if necessary. The voltage gain of the stage depends, of course, on the value of the α' for the transistors used, but at worst it is greater than 100. This is comparable with the gain by a pair of pentode valves, and there is the additional advantage that the mean signal level changes by only a few volts between input and output.

The filter formed by R_7 and C_1 reduces the ripple voltage applied to the base of X_2, and also reduces the loop gain of the system at high frequencies and prevents instability.

The apparatus has been specifically designed to work from the same high tension supplies as are commonly used condition. The effect of room temperature changes depends somewhat on the particular components used, and on the rate of temperature change, but as a working guide the change in output voltage may be taken as less than 0·05 per cent per °C change in room temperature (0·03 per cent per °F). Allowing for a 10° C difference of temperature between the transistors and the air outside the chassis, the maximum permissible room temperature is 35° C (95° F).

The circuit given in Fig. 1 can be modified fairly easily if a larger output current (2A say) is required. The essential change is the replacement of V_3, T_1, L_1 and MR_1 by components capable of handling the increased currents and voltages. The anode of V_3 is taken via T_1 to the unstabilized h.t. supply, and V_2 controls the screen grid voltage only of V_3. However, the increased size of the new components

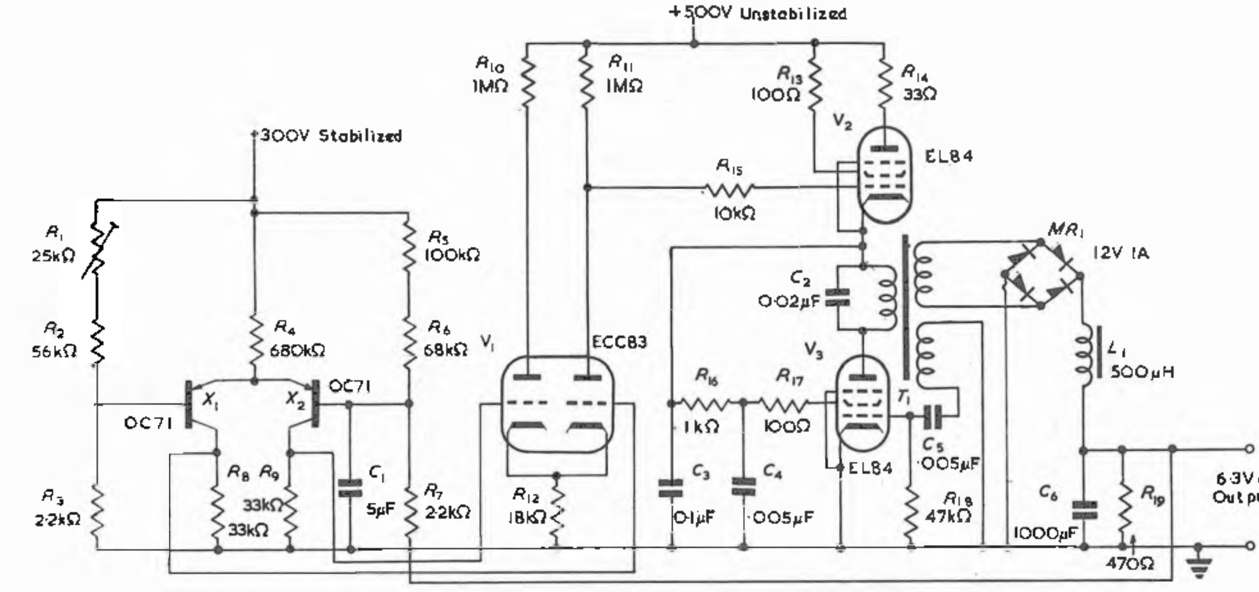

Fig. 1. The 6·3V d.c. stabilized supply

with electronic instruments. In the prototype the 500V supply is obtained from a conventional full wave rectifier system with a capacitor input filter. The stabilized 300V supply is derived from the 500V supply by means of the cascode degenerative stabilizer described by Attree[2]. The current drawn from the 500V line is less than 55mA, and from the 300V supply less than 7mA. A stabilized h.t. supply having a value other than 300V may, of course, be used; the only modifications necessary to the circuit shown in Fig. 1 are in the values of R_2, R_4 and R_6.

Performance

For output currents of between 0·8A and 50mA, the internal resistance of the 6·3V supply is less than 0·05Ω, and variations in the mains supply voltage of ± 10 per cent cause the output voltage to vary by less than ±0·05 per cent. In the extreme conditions of zero output current and a mains voltage 10 per cent greater than its nominal value, the output voltage may increase by as much as 1 per cent, but this is unlikely to be important in practice. The ripple content of the output is less than 5mV r.m.s.

When the apparatus is switched on after being idle for some time, the output voltage may slowly decrease (or increase) by as much as 1 per cent during the first half hour, but thereafter any drift is less than 0·1 per cent over an hour, provided that the load current and room temperature are reasonably constant. The initial drift is not normally serious, since the valves to which the supply is fed will take at least half an hour to reach their final operating condition.

detracts somewhat from one of the principal features of the apparatus—its compactness.

Acknowledgements

The author's thanks are due to Mr. W. M. Stratton for constructing the prototype apparatus.

Appendix

Specification for T_1

The core is made by combining the parts from two Mullard Ferroxcube pot cores type LA5 so as to form a single core having double the winding area. The air-gap is adjusted in the usual way to give the primary an inductance of 3·2mH.

Primary winding: 164 turns, 26 s.w.g. enamelled copper wire wound 82 turns to each bobbin.

Secondary winding: 11 turns, 22 s.w.g. enamelled copper wire.

Grid winding: 16 turns, 32 s.w.g. enamelled copper wire

Specification for L_1

40 turns, 20 s.w.g. enamelled copper wire wound on Mullard ferroxcube pot core type LA4. The air-gap is adjusted to give an inductance of about 500μH.

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RC Filters and Oscillators Using Junction Transistors

By N. Sohrabji*

Filters and oscillators are described in which a modified parallel-T network is used in the feedback loop of a two-stage junction transistor amplifier. The same circuit may be used either as an oscillator, or as a filter, by adjusting the gain of the amplifier. Details are given of the effects of variations of ambient temperature, supply voltage, and load impedance. Successful operation is obtained at frequencies up to 20kc/s. Over a wide range of frequency, the resonant frequency is comparatively independent of the transistor parameters.

OSCILLATORS and filters obtained by using an RC network in the feedback loop of a thermionic valve amplifier are very popular for low frequency applications, where the corresponding LC units would involve the use of bulky and expensive coils. The extension of these techniques to transistor circuits results in extremely compact units requiring little power. Transistor RC oscillators reported so far have used low impedance networks and the resonant frequency has been very much dependent on the parameters of the transistor. This article gives some of the results obtained in an effort to evolve circuits in which the following advantages would be obtained:

(a) The resonant frequency should be comparatively independent of the transistor parameters.

(b) The RC network should be worked at a high impedance, so that for a given frequency, the capacitance values would be relatively small.

(c) The operating frequency should extend beyond the audio range.

A modified parallel-T network is used as the frequency selective feedback circuit.

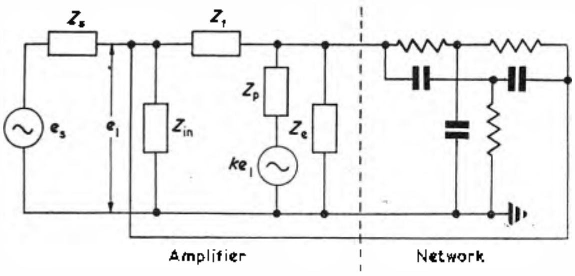

Fig. 1. The basic circuit

The Basic Circuit

This is the same for both oscillator and filter, the function being determined by the circuit conditions. A generalized equivalent circuit is shown in Fig. 1. Z_f represents the internal feedback in the amplifier, and its impedance must be kept as large as possible. It is desirable that the amplifier should be flat in the frequency range of interest; the frequency response can then be predicted from that of the network. When the amplifier gain is equal to the attenuation in the network, the circuit will oscillate at the frequency at which the total phase shift around the loop is zero. If the resonant frequency is to be dependent solely on the constants of the network, the phase-shift in the amplifier must be 180°, its output impedance must be much less than the input impedance of the network, and its input impedance much greater than the output impedance of the network. Some loading is inevitable, since the feedback voltage must appear across a finite impedance.

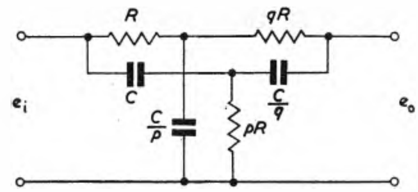

Fig. 2. The parallel-T network

The Parallel-T Network

The general case of a parallel-T network is shown in Fig. 2.

The response of the network is given by

$$e_o/e_i = \frac{(\omega/\omega_0)[pq\,(\omega/\omega_0)^2 - p(p+pq)] - j[(\omega/\omega_0)^2\,p(p+pq) - pq]}{(\omega/\omega_0)[pq\,(\omega/\omega_0)^2 - (q+p+pq+p^2+p^2q)] - j\,[(\omega/\omega_0)^2\,(p+q+pq+p^2+p^2q) - pq]} \quad \ldots\ldots\ldots\ldots (1)$$

where

$$\omega_0 = (1/RC) \quad \ldots\ldots\ldots\ldots (2)$$

when $q = 1$,

$$e_o/e_i = \frac{(\omega/\omega_0)[p(\omega/\omega_0)^2 - 2p^2] - j\,[(\omega/\omega_0)^2\,2p^2 - p]}{(\omega/\omega_0)[p(\omega/\omega_0)^2 - (2p^2+2p+1)] - j[(\omega/\omega_0)^2\,(2p^2+2p+1) - p]} \quad \ldots\ldots\ldots\ldots (3)$$

At the resonant angular frequency $\omega = \omega_0$, e_o/e_i has a minimum, real, value, given by

$$e_o/e_i = \frac{2p^2 - p}{2p^2 + p + 1} \quad \ldots\ldots\ldots\ldots (4)$$

When $p = 0{\cdot}5$, $e_o/e_i = 0$. This corresponds to the usual condition in which a parallel-T network is used. When p is less than 0·5, e_o/e_i has a finite, negative value, and hence an output is obtained which is 180° out of phase with the input.

Let $A_o = (e_i/e_o)$

Differentiating A_o with respect to p

$$dA_o/dp = \frac{1 - 4p - 4p^2}{(2p^2 - p)^2} \quad \ldots\ldots\ldots\ldots (5)$$

One value of p for which dA_o/dp is equal to zero is 0·207. Further differentiation shows that for this value of p, A_o is a minimum. Substituting $p = 0{\cdot}207$ in equation (4) shows

* Formerly Redifon Ltd.

that the minimum value of gain required for sustained oscillations is approximately 20·6dB.

For filter circuits, the gain must be adjusted to a value below that required for oscillation. The maximum value of Q which may be used is then dependent on the stability of the amplifier.

The Amplifier

A common-collector, common-emitter combination is used with RC coupling. Such a circuit has a high input impedance, a moderate output impedance, and gives a 180° phase-shift. The power supply is obtained from a 4·5V battery.

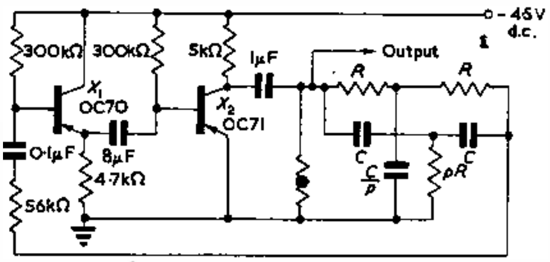

Fig. 3. The oscillator

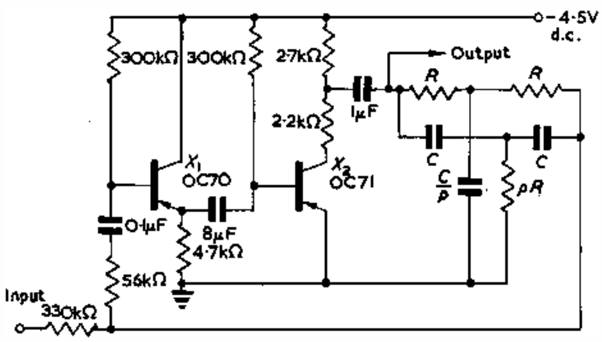

Fig. 4. The filter

The Oscillator (Fig. 3)

The input impedance of the amplifier is further increased by using a series resistor in the base circuit of transistor X_1. The parallel-T network is supplied from the collector of X_2 and feeds back into the base of X_1. A thermistor is used to limit the amplitude of oscillations and thus provide a low distortion output at constant amplitude. The output may be taken from the collector of X_2. The drain on the 4·5V battery is 0·8mA.

FREQUENCY

Oscillation is obtained at frequencies up to 20kc/s. The difference between the frequency of oscillation and the calculated resonant frequency of the network is shown in Table 1. The component values were measured to an accuracy of 1 per cent and the frequency measured to an accuracy of 0·5 per cent.

At the higher frequencies, the value of C must be increased and that of R decreased. Up to about 3kc/s, the frequency of oscillation is very close to the resonant frequency of the network. This is a major advantage of the present circuit. Oscillation has been achieved at frequencies as high as 20kc/s, but the internal reactances of the transistors have a marked effect on the frequency at this range. At very low frequencies large phase shifts occur in the coupling network, and d.c. amplifier circuits must

TABLE 1

R(kΩ)	60	60	100	40
C	1000pF	500pF	0·01μF	250pF
p	0·2	0·2	0·2	0·2
f_0, calculated, (c/s)	2650	5300	159	15900
f_0, measured, (c/s)	2514	4604	157	10640
Output voltage	1V r.m.s.	1V r.m.s.	1V r.m.s.	1V r.m.s.

be used. The thermistor is effective in keeping the voltage at various frequencies constant to within 5 per cent, and the distortion to less than 5 per cent.

STABILITY

It is important to assess the effects of variations of ambient temperature, supply voltage, and load impedance,

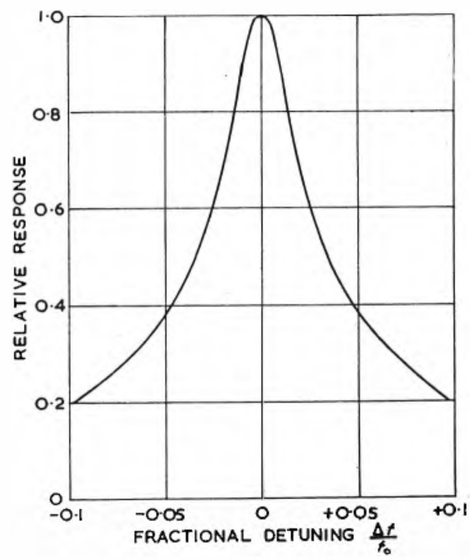

Fig. 5. Response of the filter
f_0 = 2 514c/s, Q = 25

on the frequency, waveform, and amplitude, of oscillations. The effect of temperature is shown in Table 2. The thermistor was removed for these measurements since it is inherently temperature-sensitive. It is, however, possible to reduce the effect of temperature variations on the thermistor by working it at a large fraction of its maximum power rating.

TABLE 2

Ambient temperature (°C)	20	25	30	35	40
Frequency (c/s)	2486	2479	2470	2458	2442
Output voltage (r.m.s.)	1·65	1·65	1·62	1·6	1·56
Distortion (per cent)	<10	<10	<10	<10	<10

Good stability can be achieved by using high stability resistors and capacitors, and by balancing the temperature coefficients.

Changes of supply voltage of ±10 per cent have no effect on the frequency and distortion, while the amplitude changes by ±5 per cent. Load impedances as low as 10kΩ may be used without affecting the performance of the circuit.

The Bandpass Filter

The circuit diagram is shown in Fig. 4. The same amplifier circuit is used as for the oscillator, and the overall gain is reduced to a value below that required for oscillations by tapping the collector load resistor of X_2. The Q of the filter is related to the amplifier gain and the stability of Q is therefore a function of the gain stability. The maximum value of Q which may be used, is, therefore, limited by the necessity of avoiding oscillations. A typical response curve is shown in Fig. 5. The centre frequency of the filter for a given network will be identical with that obtained for the oscillator.

Mechanical Details

Layout is generally unimportant, except at the higher frequencies. The use of high impedance networks allows comparatively small values of capacitance to be used even at low frequencies, and extremely compact circuits are possible. Resin encapsulated units have been built, which, on preliminary tests, appear to have a good performance under impact and vibration.

Conclusion

The results obtained show that the desirable features mentioned in the introduction have been realized to some extent. A very simple circuit has been used in these investigations in order that the results may be compared accurately with the predicted performance. The use of overall negative feedback and special stabilizing techniques provides greater stability against variations of ambient temperature. The development of silicon junction transistors will, no doubt, enable these circuits to approach in precision and stability, the performance of thermionic valve circuits.

Acknowledgments

The author wishes to express his thanks to the authorities of the Polytechnic, Regent Street, for providing facilities for research work.

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Variable Delay Spark Illumination for Single Shot High Speed Photography

By G. Hodgson*

A method of taking a series of high quality photographs of a reproducible phenomenon by delaying the illuminating spark by a known controllable length is described. The delay is variable from 100μsec to 1sec after the object has interrupted a light beam. The photographs are taken on a conventional camera, each photograph having an exposure time of 20μsec.

THE use of a spark as a source of illumination for single shot high speed photography is by now well established, and in cases where the phenomenon to be photographed is reproducible it is probably the best method. Normally the objects to be photographed (in this case drops of liquids) are made to interrupt a beam of light which triggers a spark when the object is in focus on a stationary camera. If a delay of controllable known length can be introduced between the object interrupting the light beam and the spark discharge, then a series of photographs can be taken, providing that the object is small compared with the field of view of the camera.

Apparatus

Fig. 1 shows a block diagram of apparatus for taking a series of photographs of a drop of liquid falling on to a liquid surface. The detector unit consists of a photomultiplier and a galvanometer lamp. The light beam from the galvanometer lamp comes to a focus 10cm from the lamp. The photomultiplier is placed 5cm behind the point of focus to allow a large patch of light to fall on to it. A drop of liquid passing through the focal point of the light beam will decrease the amount of light falling on the photomultiplier causing a pulse to be fed to the trigger unit. This unit produces a fast rising pulse which starts the delay unit, and introduces a delay variable between 100μsec and 1sec before operating the pulse unit. The function of the pulse unit is to produce a 2·5kV pulse to the triggering electrode of the spark gap.

A complete circuit of the apparatus is shown in Fig. 2. V_1 and V_2 form the trigger unit which is a bistable multivibrator[1].

A positive pulse on the grid of V_1 causes its anode voltage to fall, and the grid voltage of V_2 to drop, turning V_2 off and producing a positive pulse at its anode. A switch is provided to return the grid of V_2 to zero for resetting the multivibrator. The delay unit comprising V_5 and V_6 form a Miller linear sweep generator[2,3]. V_6 is a cathode-follower

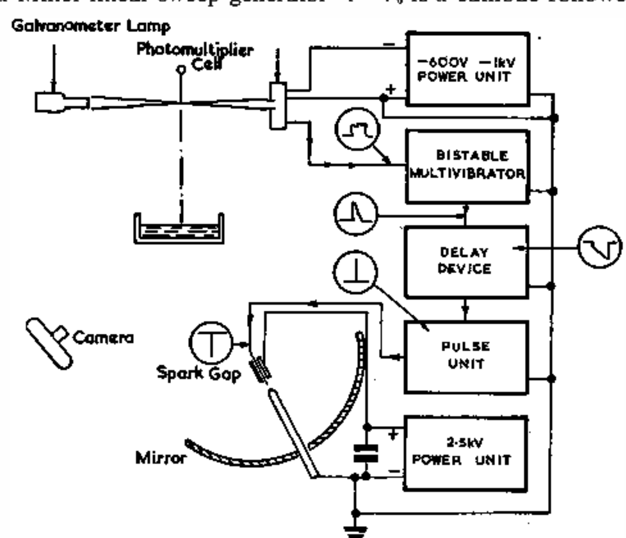

Fig. 1. Arrangement of apparatus

for the purpose of decreasing the fly-back time. The sweep valve V_5 is cut off by holding its cathode positive; the run down is initiated by the positive pulse from the trigger unit arriving on the suppressor grid. The run down or delay time is determined by C and R (in practice a range of capacitors and a variable resistor were used to vary the delay time.) The output from the cathode of V_6 is fed to the pulse unit V_8, this is a CV73 (11E3) pulse tetrode which is cut off by holding the cathode 50V positive. Only a positive pulse is allowed to arrive at the grid of V_8, V_7 by-

* *Research Laboratory for the Physics and Chemistry of Surfaces, Department of Physical Chemistry, Cambridge.*

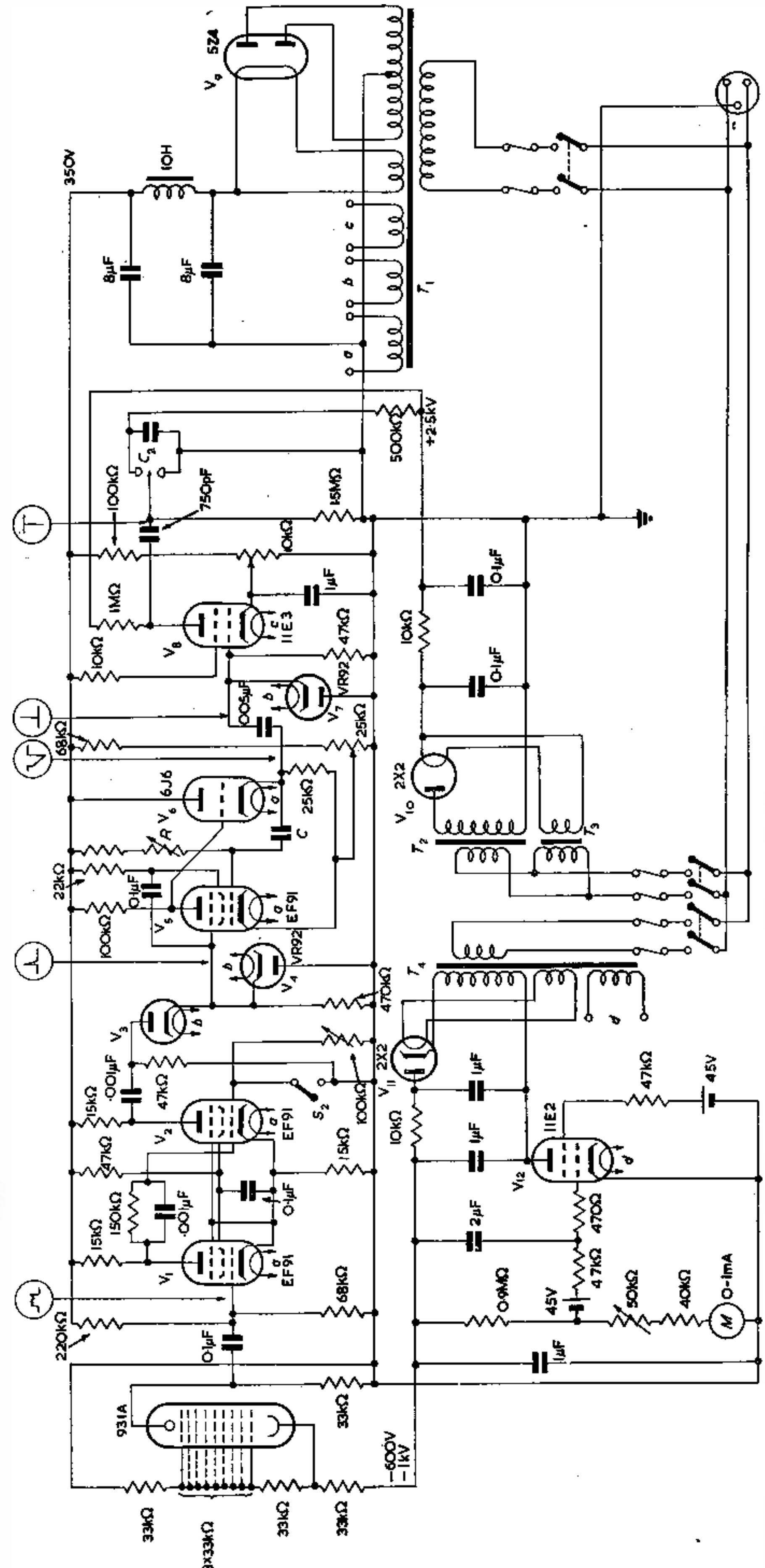

Fig. 2. The complete circuit

passing all negative pulses. The anode of V_8 is at 2·5kV, and when the valve is turned full on by a positive pulse arriving at the control grid from V_6 a pulse of about 2kV is fed to the triggering electrode of the spark gap through a 750pF ceramic cup capacitor.

The spark gap is a 3 electrode type[4,5] and was built across the terminals of an 8μF 4kV capacitor.

The earthy electrode has a triggering electrode passing through its centre and insulated from it with a ceramic tube. Stainless steel was used for the electrode tips which are spaced about 4mm apart.

A large elliptical mirror from a projection arc lamp was used to focus the spark on to the area in which the drops were to be photographed.

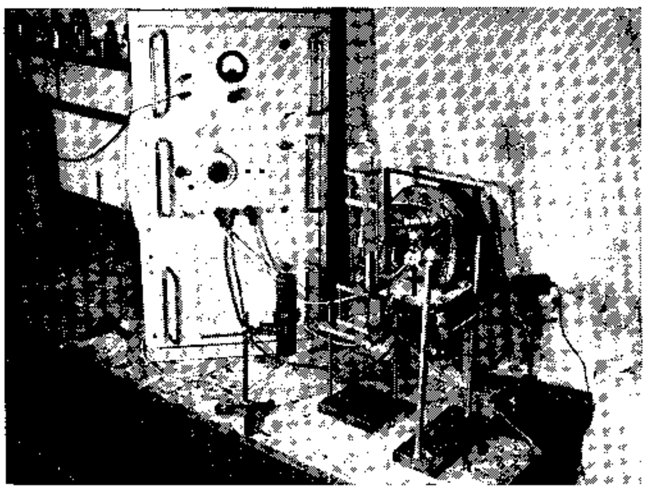

Fig. 3. The complete instrument

Power Supplies

Three power units were required for the apparatus. A 600V to 1kV negative supply is required for the photomultiplier, this is semi-stabilized against mains and h.t. variation. A 50kΩ potentiometer in the grid circuit of the CV276 (11E2) stabilizer valve provides the required control over output voltage, a 0 to 1mA meter indicates the output voltage and is scaled 0 to 1 000V. The trigger unit and delay unit require 350V d.c. which was obtained from a conventional power unit as shown in Fig. 2. The 8μF storage capacitor and the pulse output valve require 2·5kV d.c. This was obtained from a half-wave rectifier power unit as shown in the circuit diagram. The storage capacitor is charged through 500kΩ resistor which limits the current drawn from the power unit when the spark gap discharges.

The apparatus was assembled as shown in the photograph Fig. 3.

A Contax MKII camera with a Sonnar $f/1 \cdot 5$ lens was used to photograph the drops on Panatomic X film, the effective numerical aperture being $f/22$. As the shutter of the camera had to be left open it was essential to operate the equipment in a darkened room (but this did not prove difficult in practice). The exposure time is determined by the time taken for the storage capacitors to discharge through the spark gap, in this case 20μsec.

Fig. 4 shows 3 photographs taken from a series of 30. These are of a drop of water falling from a height of 1·3cm on to a water surface, and show the impact. The water had added to it a small amount of synthetic detergent, Teepol.

Conclusion

It has been shown that by using a spark discharge as a source of illumination and varying the time between an object passing through a beam of light and the spark discharge, a series of good quality photographs can be taken, each photograph having an exposure time of 20μsec. It must be stressed, however, that the phenomena to be photographed must be reproducible.

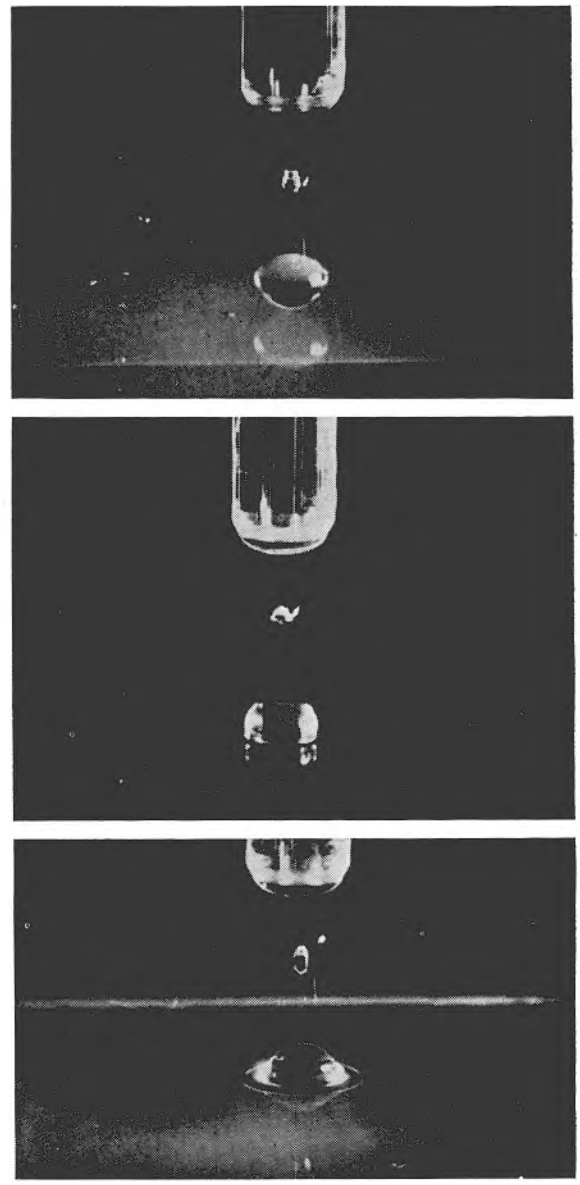

Fig. 4. Water drops falling on to a water surface

The apparatus which has been described proved adequate for the purpose for which it was designed, the accuracy of the delay unit could, however, be improved if more precise measurements were required, by the addition of catching diodes in the delay circuit.

Acknowledgments

The author would like to thank Mr. J. O'Callaghan who is making a study of liquid drops for initiating this work, and to Vauxhall Motors for a grant to the Laboratory.

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LETTERS TO THE EDITOR

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

The Adjustment of Linear Logarithmic Function Generators

DEAR SIR,—A direct-coupled function generator may be tested by feeding in known voltages and measuring and plotting the output voltages. This method is, however, very tedious and it does not lead to a rapid determination of the optimum settings of the various bias adjustments. In a diode function generator these inevitably require some empirical trimming since the characteristics of individual diodes vary appreciably.

With a linear-logarithmic function generator, matters are simplified by applying at the input a voltage which decays exponentially towards zero. When the function generator is correctly adjusted, this input should result in an output varying in a linear manner with time. If the exponential input signal is made to recur at a convenient frequency, the output may be observed on an oscilloscope and the effect of each bias adjustment may then readily be studied.

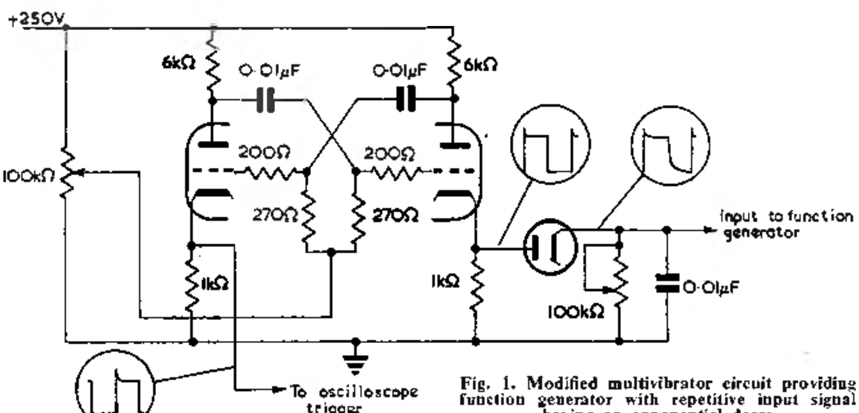


Fig. 1. Modified multivibrator circuit providing function generator with repetitive input signal having an exponential decay

The necessary input signal may be provided by the circuit shown in Fig. 1. This comprises a form of multivibrator from which a substantially rectangular output waveform is obtained having an unimportant over-shoot on the positive going step. At each cycle, the positive going step on the waveform rapidly charges the capacitor *C* through a diode so that the input to the function generator is raised to about −20V. Half a cycle later, the multivibrator output falls abruptly to zero. This cuts off the diode so that the capacitor *C* discharges through resistor *R* according to an exponential law.

With an ideal logarithmic function generator, such an input would produce an output changing in a linear manner with time. Fig. 2 shows a typical result obtained from a simple four-diode function generator. The plateau at the beginning of the trace corresponds to the condition in which *C* is fully charged. Some curvature is clearly visible at the foot of the trace, corresponding to inputs below the range of logarithmic transfer. In the central part of the characteristic, however, the departures from linearity are barely discernible and in a function generator using many more diodes it becomes quite impracticable to adjust individual bias controls by observation of such a display. Even a simple circuit may not be adjusted to the optimum condition since non-linearities in the oscilloscope amplifier, time-base and c.r.t. distort an otherwise linear diagonal trace. This method of display is therefore not satisfactory where output errors must be restricted to less than about 5 per cent of the total deflexion.

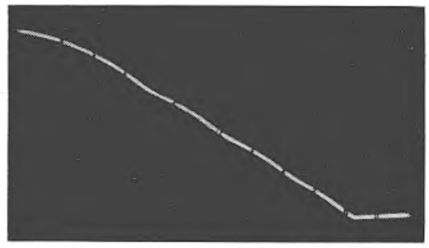


Fig. 2. Output from simple 4-diode function generator with input provided by circuit of Fig. 1. Ideally this trace would be a straight line

By passing the output signal through a differentiating stage, an oscilloscope trace may be obtained representing the gradient of the output rather than the output itself. With a true exponential input to the function generator, therefore, the oscilloscope display should consist of a horizontal straight line over the operating range of the circuit. When a reasonable approximation to this has been obtained, the oscilloscope gain may be increased until it is seen that this line consists in fact of a series of humps. This is clearly evident in Fig. 3 which shows the differentiated and amplified signal obtained from the output shown in Fig. 2. The humps in the oscilloscope display arise where the generated function departs locally from the ideal logarithmic function. When only a small number of diodes is used, the departures from the ideal will necessarily be considerable. In the case of a function generator using 16 diodes, however, the departures from a true logarithmic response can be restricted to less than 1 per cent over a range of three decades. In Fig. 4 the amplification has been increased by a factor of five in order to reveal the much smaller irregularities in the differentiated output. These irregularities may always be made visible, however, by sufficiently increasing the amplification and it is then a simple matter to adjust even a large number of bias controls to their optimum settings.

Fig. 3. Differentiated output from 4-diode function generator. The changes in slope barely visible in Fig. 2 here become evident as pronounced humps. Ideally this trace would be a horizontal straight line

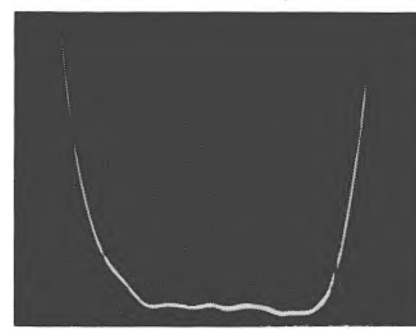


Fig. 4. Differential output from 16-diode function generator. This represents a much better approximation to the ideal horizontal straight line despite the use of an oscilloscope sensitivity higher than that used for Fig. 3

In principle, this method of testing function generators is applicable also to functions other than logarithmic provided an accurate and repetitive inverse function can be applied to the input. It happens that this is most readily done with a logarithmic function generator. Provided the observed part of the display does not correspond to an overload condition of the amplifier, this method of testing eliminates all errors due to non-linearity of the oscilloscope amplifier and time-base and geometrical distortions occurring in the cathode-ray tube.

Yours faithfully,

D. M. NEALE,
Physics Research Laboratory,
Ilford Limited.

BOOK REVIEWS

Atomkraft

By Friedrich Münzinger. 224 pp. 171 figs. Large 8vo. 2nd Edition. Springer-Verlag. Berlin/Göttingen/Heidelberg. 1957. Price DM 29·40.

IT certainly happens not very frequently that the first edition of a technical book and its reprint is sold out after two years and that the necessity arises to enlarge and mainly rewrite it after such short a time. But this is less astonishing if the subject of the book is in such rapid development as the peaceful uses of atomic energy. The first edition was reviewed in the April 1956 issue of ELECTRONIC ENGINEERING. Since then the 90MW Calder Hall station and a 5MW station in Russia have come into actual operation, the 1955 Atom-Congress in Geneva has finished publishing its proceedings, the 5th World-Power Conference has taken place in Vienna 1956 and a vast amount of new material has appeared in the technical press of all countries concerned.

While the first edition was a slender volume of 94 pages the second one has been enlarged to 224 pages. This made it possible for example to devote now about 45 pages to the fundamentals of nuclear physics while these were dealt with on 2½ pages in the previous edition, and in a similar though not quite so striking manner the other chapters have been enlarged.

The book is now subdivided into three main parts, a theoretical, a technical and an economic one. In the first part the physical principles, the cooling of the reactors and the constructional and active materials are discussed. The second part contains the description of a large number of reactor designs which, however, are mostly in the project state only, the heat engines used in the atomic power plant and the construction of complete plants. In the third part the competitiveness of nuclear power stations is discussed, a special section is devoted to developments in Germany and a prediction is made about the second industrial revolution caused by atom power and automation.

In an appendix the most important concepts are briefly explained, some useful tables are given and a not very elaborate bibliography, a name and a subject matter index conclude the book which is a welcome addition to the literature about this so important subject.

One special point must be mentioned. A footnote on page 89 states: "Immediately before the printing of this book was finished I received a communication that Calder Hall, which had been officially put into commission on the 17 October 1956, had to be shut down at the end of November because welding seams of the reactor or the steel container for the heat exchangers had shown leaks. Besides it was said that obnoxious radioactive phenomena had appeared: how far these were connected with the leaks was not evident from the communication. Furthermore it appears that the sodium-cooled reactor of the submarine *Seawolf* had to be reconstructed due to serious corrosions. Even if both communications should be true one had to expect such children's diseases, disagreeable as they may be, for the reasons given in this book. It is therefore only to be hoped that one will soon succeed in overcoming them."

As far as Calder Hall is concerned it may be said that the author's hope has been fulfilled. The shutting down in November was scheduled beforehand for a routine inspection. The very slight leaks could be mended very quickly and also deformations of the fuel elements, predicted by the Harwell specialists, could be set right within the shortest time. The plant is in full operation, the first reactor since middle of November 1956 and the second since February 1957, and it may be added that only recently Sir Christopher Hinton stated that the putting into commission and operation of the plant was rather a "dull" affair as so little happened in comparison with conventional plant.

R. NEUMANN

Automation in Business and Industry

By Eugene M. Grabbe. 611 pp. 276 figs. Demy 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd. 1957. Price 80s.

THIS book is composed of a collection of lectures and represents a habit which is becoming too prevalent at the present moment: in this case, however, there is a difference. The lectures are a deliberately planned series given in the University of California. The course was held in 1955 and was designed to provide up-to-date information on developments in automation and to describe automation system application. The lecturers were nationally prominent engineers and scientists from industry, business and the universities.

An introduction to the series is given by the Professor of Engineering at California University, in which he surveys the social implications and historical development of a science. There then follow eighteen lectures which show how the fields of feedback control theory, instrumentation, analogue and digital computation and data processing are now becoming integrated as automation is applied on a broad scale to control systems that cover the range from top management down to industrial machines.

There are separate lectures which specialize on industrial process controls, on digital computers, and on analogue computers, thereby giving a clear idea of the differences in the requirements and the functions of the equipment normally covered by these terms. In order to use the equipment in complex systems, it is necessary to have further devices which can convert from one type of presentation to another. There are, therefore, lectures describing these devices under headings such as Analogue-to-Digital Conversion Units.

The first half of the book contains lectures on the general theory of all these systems: the second half is composed of lectures which give an immense amount of factual information on applications in different fields. An example of this is a detailed lecture on the automatic control of flight of aircraft and guided missiles, which is followed by one on the automatic production of electronic equipment which is of very definite interest to most electronic engineers.

Automatic process control equipment in the petroleum and chemical industries is given a special lecture with again, details of equipment employed, while other lectures describe analogue computers used for industrial control systems.

The book finishes with lectures on the control of machine tools by a combination of servo mechanisms and digital equipments, as well as three general lectures on automation, delving into both the general economics and the future of such equipment.

The book as a whole is certainly an invaluable document for any electronic engineer interested in the general field now covered by the term "Automation." It covers the field more adequately than any other book on the market at the moment: although it is rather expensive, it is very nicely produced.

J. R. BOUNDY

Dry-Battery Receivers with Miniature Valves

By E. Rodenhuis. 242 pp. 248 figs. Demy 8vo. Philips Technical Library, Holland. Cleaver-Hume Press Ltd, London, 1957. Price 32s. 6d.

THIS book deals with the home construction of modern battery receivers and will be of considerable interest to the service engineer and amateur.

One chapter has been devoted exclusively to the question of current supply. The following pages also contain complete valve data and characteristics of miniature valves for dry-battery operation, and eight suitable circuits are given, including a.m./f.m. designs and push-pull output stages.

F.M. Radio Servicing Handbook

By Gordon J. King. 192 pp. 100 figs. Demy 8vo. Odhams Press Ltd. 1957. Price 25s.

THIS handbook has been written with the aim of providing this theoretical and practical knowledge in a form helpful to all concerned with service work. The book is intended not only for the professional service engineer but also for the amateur enthusiast interested in the construction of f.m. equipment and for radio students. The style is straightforward and, as far as possible, non-mathematical.

Progress in Semiconductors Volume 2

Edited by A. F. Gibson, P. Aigrain and R. E. Burgess. 277 pp. 70 figs. Demy 8vo. Heywood & Co. Ltd. 1957. Price 63s.

THIS is the second annual volume of review articles on selected topics, by specialists in the semiconductor field. Eight subjects are treated in this volume. The first article — *Semiconductor Alloys* — deals with disordered crystals of the solid substitutional type. It is mainly theoretical, but where comparisons with experiment are possible, emphasis is placed on the binary Ge-Si system, because the basic properties of the constituents are reasonably-well understood. The subsequent article — *Properties of III-V Compound Semiconductors*—is a well-balanced presentation of the preparation, properties and applications of the nine ordered compounds, which may be made by combining in stoichiometric proportions one element from each of the groups Al, Ga, In and P, As, Sb. In *Radiation Effects in Semiconductors* the main concern is the Frenkel type of defect (interstitial atom plus vacancy), with particular reference to experimental results for germanium and silicon. There are errors in the captions to figures (6) and (12) of this article. A clear physical picture of the processes of recombination and trapping is given in terms of a general model in *Lifetime of Free Electrons and Holes in Solids*. The special cases of phosphors, transistor materials and insulators are subsequently considered. The advantages of horizontal growing are discussed in *The Production of High-Quality Germanium Single Crystals*. A low density of dislocations is a particular feature of this method. Comprehensive coverage of the effects of a wide range of both substitutional and interstitial chemical impurities on the properties of germanium is presented in *Impurities in Germanium*. The article *High Electric Field Effects in Semiconductors* which is illustrated by experimental results relating to germanium and silicon, draws attention to the discrepancy between the observed and theoretically-expected energy loss of high-energy carriers to the lattice vibrations. Consideration is given to the effects of avalanche multiplication of carriers on device properties. Finally the relative merits of the divergent *Thories of Electroluminescence*, which are based on the impact-acceleration mechanism, are considered.

The book is a valuable addition to semiconductor literature. The presentation is such as to appeal most readily to the research physicist. If more emphasis were placed on the application of theories and basic properties to the performance of devices, wherever possible, the series would no doubt reach a wider audience. Further, as an aid to the scientist who does not even have the time to read review articles fully, the uniform inclusion of a "Conclusions" section would be a great help.

F. J. HYDE

Polythene: The Technology and Uses of Ethylene Polymers

By A. Renfrew and Phillip Morgan. 642 pp. 75 plates. Royal 8vo. Iliffe & Sons Ltd. 1957. Price 126s.

THIS book has been compiled under the joint general editorship of Mr. A. Renfrew, Director of the I.C.I. Plastics Division and Mr. Phillip Morgan, Editor of British Plastics. Thirty-eight specialists, drawn from Great Britain and the United States, have contributed to its 32 chapters.

The chapters are grouped under three main headings: Manufacture and Properties; Processing Techniques; and Applications. The last section contains chapters on future trends.

The book is intended to serve as an authoritative text for technologists not only in the plastics industry but in those many other industries where polythene is used in considerable quantities.

Transistor A.F. Amplifiers

By D. D. Jones and R. A. Hilbourne. 152 pp. 94 figs. Demy 8vo. Iliffe & Sons Ltd. 1957. Price 21s.

THIS book deals with the design of transistor audio-frequency amplifiers, and gives the circuit and design details of a versatile range of amplifiers, including those for h.f. reproduction and public address systems with undistorted outputs up to 20W.

ELECTRONIC EQUIPMENT

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

VALVE PROTECTING TIMER

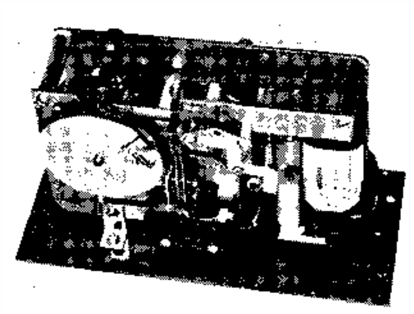
(Illustrated below)

Venner Ltd, Kingston By-Pass, New Malden, Surrey

VENNER Limited have introduced a delayed reset timer, type TDR/S/P635, designed for the protection of thermionic valves in transmitting, receiving and control equipment.

After the initial switch on, the timer operates as a standard delay relay. This ensures that the h.t. voltage is not connected until the filaments are warmed up. Upon switching off, the timer resets to zero at a delayed rate controlled by an escapement. This delayed reset feature prevents the full initial delay period having to be completed after a supply interruption of a shorter duration than the normal warm up time.

The reset time can be either the same as the initial delay period or set for some proportion of this period, e.g. a 5 minute delay time, 2½ minute reset



time. The unit can therefore be calibrated for the cooling characteristic of the equipment under control. A further feature is that the time contacts can be arranged to stay closed during very brief supply interruptions of less than say 10 seconds duration, opening only when the relay is de-energized for longer than this period.

The timer is fitted in a steel container measuring 8in by 4¾in by 3½in with connexions brought out to sealed terminals. Its operating temperature range is −40°C. to −70°C.

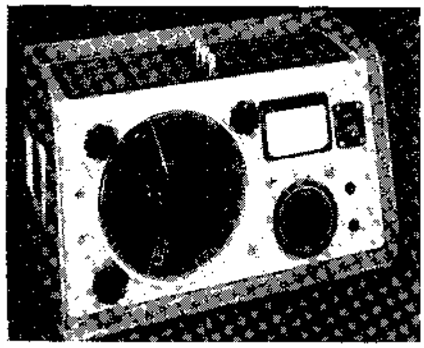
UNIVERSAL BRIDGE

(Illustrated above right)

Marconi Instruments Ltd, St. Albans, Hertfordshire

MARCONI Instruments announce the introduction of a new version of the well-known Universal Bridge TF 868/1. While retaining all the advantages of the TF 868/1, the new instrument—Type TF 868A—incorporates a bridge-volts monitoring system.

For inductance or capacitance measurements, the measuring bridge is energized by an a.c. voltage derived from a valve oscillator. If desired, this voltage can be set to a definite required value. For



this purpose there is a continuously variable *L* & *C* bridge voltage control, an 0- to 10V auxiliary scale on the panel meter, and a read *L* & *C* bridge voltage switch. The inclusion of this new facilty is especially useful when measurements are being made on components such as iron-cored coils whose inductance is a function of exciting current.

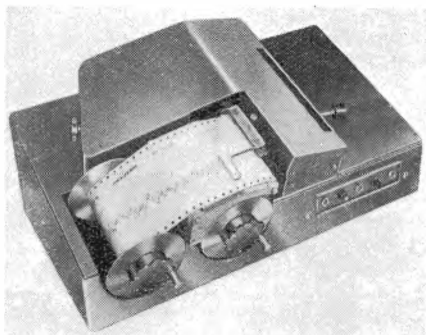
The Marconi TF 868A Universal Bridge measures values of inductance from 1μH to 100H, capacitance from 1pF to 100μF and resistance from 0·Ω to 10MΩ. Inductance or capacitance measurements can be made at either 1 or 10kc/s; resistance measurements are made at d.c. The instrument is a complete and comprehensive measurement system. Bridge networks, precision standards, a.c. and d.c. bridge supplies, an amplifier-detector for a.c. balance determination, a centre-zero galvanometer for d.c.—all are combined in one mains-operated functionally-designed instrument

TRACE READING EQUIPMENT

(Illustrated below)

Southern Instruments Ltd, Frimby Road, Camberley, Surrey

THIS is an entirely new trace reading equipment for the analysis and reduction of analogue trace records to digital form. Film or paper charts up to six inches in width are dealt with, the equipment incorporating such desirable facilities as linear or non-linear calibration and zero adjustment over the full



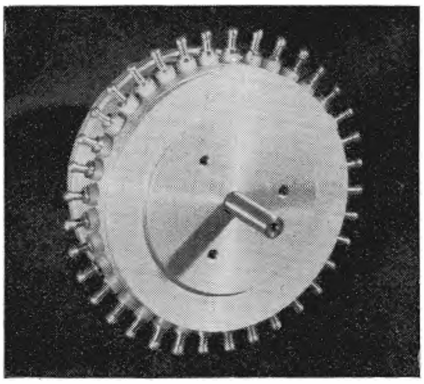
record width. In addition to producing a permanent digital record by typewriter, tape or card punch a "quick look" facility is provided in the form of a 4 digit visual display. Existing users of the earlier K1015 trace reader with fixed scale read out may now easily convert to the new system and enjoy the added facilities.

MULTI-TAP LINEAR POTENTIOMETER

(Illustrated below)

The General Electric Co. Ltd, Magnet House, Kingsway, London, W.C.2

A NEW precision linear potentiometer which can be readily adapted to provide a wide range of non-linear functions has been developed by Salford Electrical Instruments Ltd, an associate of the G.E.C.



Primarily a linear potentiometer with an electrical angle of 340°, the winding is tapped at 10° intervals and the connexions taken to turret-type terminals spaced around the potentiometer body. It is particularly suitable for producing non-linear functions by connecting shunt resistors across sections of the winding. The large number of taps enables a wide range of functions to be accurately reproduced.

In equipment such as flight simulators where a large number of different functions is required, the replacement of faulty potentiometers of the shaped card type becomes a problem. Each of the many different types of functional potentiometers required must always be available. The new multi-tap potentiometer, however, can be carried in stock as a standard unit and adapted to suit the particular needs as required.

An additional advantage of the use of the multi-tap potentiometer arises in the laboratory where non-linear functions can be rapidly produced and modified at will.

Of robust construction and low starting torque the potentiometer can be supplied in a range of resistance values from 1kΩ to 100kΩ.

PORTABLE TRANSISTORIZED COUNTER

(Illustrated above)

P.A.M. Ltd, Merrow, Guildford, Surrey

THIS battery operated portable transistorized counter has been designed for counting small production quantities at a maximum rate of four per second. It has a built-in photo-electric cell, or alternative remote photo-electric cell probe. The battery life gives 250 000 continuous counting operations. The casting is strongly constructed of hammered finished steel. The dimensions are: 4¼in by 3¾in by 1⅞in.

PROCESS TIMER

(Illustrated below)

D. Robinson & Co., 58 Oaks Avenue, Worcester Park, Surrey

A NEW self-resetting synchronous timer has just been added to the Rodene range. It is designed for through panel mounting and, except for a small steadying screw, is one-hole fixing.

It employs the instant-start, self-clutching Rodene timer motor which makes possible the simple method of operation shown in the adjoining cut-away view of the new unit.

The spring marked E normally holds the arm A against the stop C which is positioned by the control knob. The instant the timer is switched on the arm begins moving counter-clockwise, the gearing in the motor being such that it takes the time indicated on the dial to reach the "O" position where it depresses the actuator D of a heavy duty micro-switch.

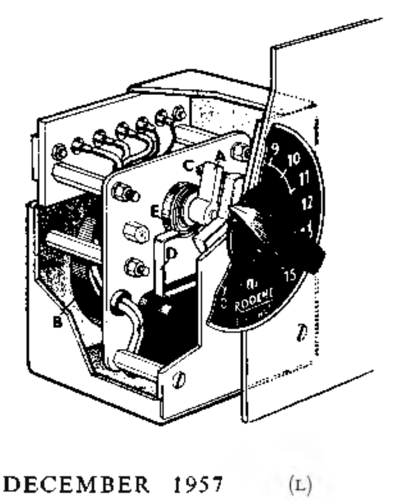
If the motor is now left energized it will come to a standstill, but will hold the switch operated. When the motor is switched off its integral clutch opens allowing the spring to reset the switch immediately and restore the arm to its normal position against the stop ready for the next cycle.

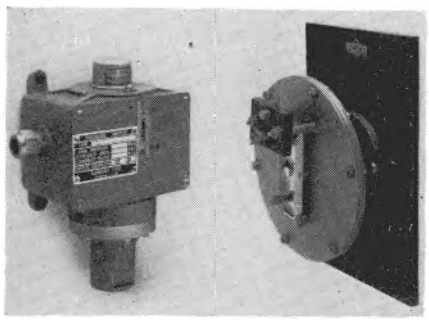
There are four standard models, providing ranges of 0 to 15sec; 0 to 60 sec; 0 to 5min; and 0 to 10min. Due to the low stalled current and torque the first two models can be left 'timed out' with the switch actuated for as long as required.

The two larger period models each have a built-in relay arranged to reset the mechanism at the end of the timed period and to hold a contact operated during the 'timed out' period thus providing effectively the same action.

Behind-panel dimensions of the 'seconds' timers are only 3½in high by 3in wide by 3in deep. The 'minutes' models are 3½in by 3in by 5in deep. The output switch has changeover action and is rated at 3A at 450V a.c. and 5A at 250V a.c. Motors are wound for most standard voltages and frequencies.

Accuracy is high as the special motor always reaches synchronous speed in about two cycles.





PRESSURE SWITCHES

(Illustrated above)

Londex Ltd, Anerley Works, 207 Anerley Road, London, S.E.20

THERE are two new additions to the range of Londex pressure switches; these are the types POS and PS/P.

The POS switch operates on extremely low pressures, it can be used for a variety of applications—if coupled to a rubber tube it can be used as a limit switch, for vehicle counting, as an alarm device to give warning of persons entering or leaving premises, or for door opening control.

This switch will operate on pressures between ½in and 3in w.g., with a reset differential of ¼in w.g. The single-pole changeover contacts are rated at 2A at 440V a.c.

The PS/P switch is designed for operation on oil hydraulic systems, the build-up of the switch being similar to the other standard switches but utilizing a piston as the pressure sensitive element instead of bellows.

This type of switch is available for operation in the range of 50 to 5 000lb/in² with a reset differential between 15 and 300lb/in².

The type PS/P is illustrated on the left and the POS on the right.

MINIATURE INSULATORS AND TERMINALS

(Illustrated above)

Horwin Engineers Ltd., 101-105 Nibthwaite Road, Harrow, Middlesex.

THE W.8003 is a miniature stand-off insulator. Standing only ¾in high it is moulded in Alkyd and is for 1kV working. Alternative bases with either a single or double top, solder dipped for easy soldering, can be supplied.

The 'Forway' terminal is in I.C.I. Alkathene and available in a range of five distinctive colours. It will accept standard 'Banana' plugs, taper pin connectors, spade lugs, solder tags, or bare wires and is suitable for currents up to 10A.

UNIT CONSTRUCTION RESISTORS

Radiospares Ltd, Maple Street, London, W.1

THESE unit construction resistors are designed primarily for use as mains dropping resistors in radio and television receivers, the units being constructed so that any required number can be mounted together on a suitable length of 2 B.A. studding. The windings are totally enclosed in a round ceramic body, cement sealed. The units are available in 10W and 20W sizes in a range of values between 7Ω and 470Ω.

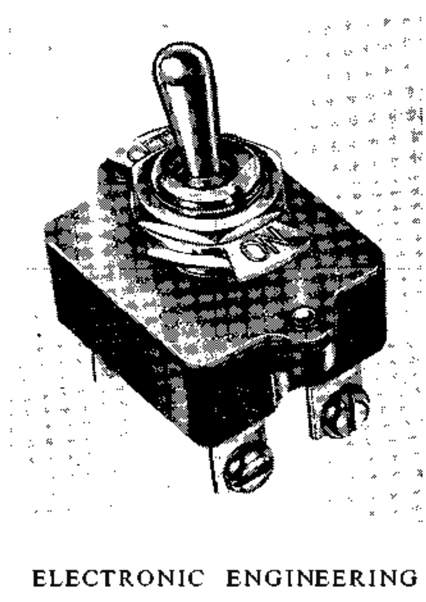
MINIATURE TOGGLE SWITCH

(Illustrated below)

Arcolectric (Switches) Ltd, Central Avenue, West Molesey, Surrey

THIS switch type S.254, is rated 10A, 250V a.c. only. It is double pole on-off. Features of the design are long life—it has been type tested on its full rated load to a quarter of a million operations—small size—it is an exceptionally compact double pole 10A switch with readily accessible screw terminals.

The switch has a positive snap action and is designed on the micro gap principle with fine silver contacts.



Short News Items

The Sixth Annual Convention of the Scientific Instrument Manufacturers' Association has maintained the usual high attendance, when over 240 people assembled at Eastbourne representing a comprehensive cross-section of the scientific instrument industry. The Convention, "Enterprise 1958," was opened on 25 October by the Marquess of Exeter, Chairman of A. C. Cossor Ltd. The theme of the Convention looked to the future, particularly to the context of the British industry in the European Free Trade Area. Four discussion panels covered the following subjects:— *European Free Trade—A Challenge,* Chairman: Mr. J. E. C. Bailey, C.B.E. (Baird & Tatlock (London) Ltd); *Presentation and Publicity,* Chairman: Mr. A. G. Peacock (Mervyn Instruments); *Methods and Materials for Tomorrow,* Chairman: Mr. M. G. D. Curties (C. Baker of Holborn Ltd); *Nucleonics and The Nuclear Industry—Instruments for Reactors and Safety,* Chairman: Mr. H. A. Luss (Isotope Developments Ltd).

A new Research Establishment for the Research Division of Powers Samas Accounting Machines Ltd has recently been opened by Viscount Knollys. Situated at Whyteleafe, Surrey, the new buildings cover 50 000ft² on a site of three acres, and accommodate the laboratories, workshops, drawing office and administrative offices. The Company's Research Division comprises five technological branches: Research and Development, Design, Workshops, Product Improvement, Automatic Recording and Control. These divisions are served by a Technical Information Services Section covering library, patents and technical writing. Among the latest developments of the Company are the electronic computer (Empe), the Programme Controlled Computer (P.C.C.) and the "Samastronic" tabulator. The latter employs an oscillating stylus and is capable of printing characters at a speed of 42 000 characters a minute.

The scope of the facilities of the Mullard Research Laboratories at Salfords, Surrey, has been increased by the opening of extensive new buildings. These were made necessary by the growing demands being made on the Laboratories, not only by the various divisions of the Company, but also by Government Departments. The main addition is a two-storey laboratory building flanked by a four-storey office block, providing a total of 28 000ft² of floor space. Of this, the Laboratories proper occupy some 18 000ft² and the remainder is in use as offices and a large technical library. Total staff now numbers about 700, of whom 175 are graduate physicists or engineers.

The BBC has awarded two research scholarships, valued at £435 per annum, to university graduates in Electrical Engineering, giving them the opportunity to work for a higher degree at any university in the United Kingdom. The first is to Mr. J. B. Izatt, who graduated at Aberdeen University with a 1st Class Honours Degree in Electrical Engineering. Aberdeen University engineering students attend the electrical engineering laboratories at Robert Gordon's Technical College and Mr. Izatt will conduct his researches there under the guidance of Dr. E. Wilkinson and Mr. A. M. Hardie. The second scholarship has been awarded to Mr. W. A. G. Voss who graduated at Queen Mary College, University of London with a 2nd Class Honours Degree in Electrical Engineering. Mr. Voss will conduct his researches in the Department of Electrical Engineering at Queen Mary College, London, under the guidance of Professor M. W. Humphrey Davies. The scholarships, which are awarded annually are for two years in the first instance with the possibility of extension in suitable cases, if necessary. The scholarships are limited to male British subjects normally resident in the United Kingdom and the only condition applying to the subject for research is that it must be in those fields of telecommunications or physics which have an application to sound or television broadcasting.

The Radio Industry Council has announced that the 25th National Radio and Television Exhibition will be held at Earls Court, London, from 27 August to 6 September, 1958.

The Radio Communication and Electronic Engineering Association, twenty-one of whose members exhibited in the first Instruments, Electronics and Automation Exhibition last May, has announced that it has become one of the sponsoring associations for the coming exhibition to be held at Olympia from 16-25 April next. Other sponsoring bodies are the British Electrical and Allied Manufacturers' Association, the British Industrial Measuring and Control Apparatus Manufacturers' Association, the British Lampblown Scientific Glassware Manufacturers' Association, the Drawing Office Material Manufacturers' and Dealers' Association, the Scientific Instrument Manufacturers' Association of Great Britain.

Three new chairs have been created at the Royal Military College of Science, Shrivenham, and the following appointments have been made. Dr. A. Charlesby, Professor of Physics; Mr. A. D. S. Carter, Professor of Mechanical Engineering; Mr. Alan Lee, Professor of Electrical Engineering (Radar and Telecommunications). Three new members of the Advisory Council, which advises the War Office on matters of policy connected with the College have been appointed in place of members whose term of office has expired. They are: Professor J. Diamond of Manchester University; Professor M. H. L. Pryce, F.R.S., of the H. O. Wills Physical Laboratory, Bristol; and Dr. R. P. Bell, F.R.S., of Balliol College, Oxford.

Scandinavia's biggest electronic computer and data processing machine, the Atvidaberg Facit EDB, has just been completed. It has been built by a team of mathematicians and engineers headed by Mr. Erik Stemme, who before joining the Atvidaberg electronics department last year was also responsible for the construction of the Swedish Government's BESK electronic computer. An improved version of BESK, the Facit EDB can perform 23 000 additions or 3 600 multiplications of 12-digit decimal numbers per second. Programmes fed into the machines are stored in a fast ferrite memory with a capacity of 2 048 12-digit decimal numbers. In addition, there is a magnetic drum memory with a storage capacity of 8 192 12-digit decimal numbers. The unit has over 2 000 electronic valves plus transistors. The power consumption is 17kW. The Facit EDB will form the nucleus of Atvidaberg's data processing centre, the services of which will be available for Swedish business enterprises, scientific institutions, governmental and municipal authorities, etc, both on a contract basis and for temporary assignments.

A further Elliott G-PAC General Purpose Analogue Computer has been delivered, this time to the Northampton College of Advanced Technology in London, where it is to be operated by the Mathematics Department, mainly for the training of students in the use of Analogue computing techniques in applied mathematics. Other departments of the College will also use the Computer for servo analyses and similar problems. In addition to the G-PAC certain ancillary equipment has been ordered by the College, including one of the latest Elliott High-Performance Servo-multiplier Units and various Non-Linear Elements.

The new Electronics Department of Bruce Peebles & Co Ltd at 37 Inverleith Place, Edinburgh, was recently officially opened by The Rt. Hon. Thomas Johnson, C.H., LL.D., Chairman of the North of Scotland Hydro-Electric Board and Former Secretary of State for Scotland.

The Automatic Telephone & Electric Co Ltd have been awarded two important contracts in connexion with the extensive trunk telephone modernization programme now being undertaken by the New Zealand Posts and Telegraphs Department. The first calls for the provision of a coaxial carrier system, capable of carrying many hundreds of telephone circuits, to link the important North Island centres of Wellington and Palmerston North. Seventeen unattended repeater stations will be employed along the 95 mile route separating the two cities, which in the past have been served by carrier systems on overhead wires. At present A.T.E. are engaged in the manufacture and installation of a similar coaxial link between Auckland and Hamilton, which when completed will be the first of its kind in the Dominion. Both links will cater initially for between 240 and 300 high grade telephone circuits, with an ultimate capacity of 960 circuits. The second order is for the Company's latest channelling equipment for use in conjunction with a microwave radio link between Palmerston North and Hamilton.

The Ministry of Transport and Civil Aviation have announced that a radar observer course leading to the issue of a Certificate of Proficiency as Radar Observer in a form approved by the Ministry, is now available at the School of Navigation, Royal College of Science and Technology, Glasgow.

The journals Optica Acta and Acta Electronica have undertaken the joint publication of the full proceedings (invited lectures and communications, mostly in English) of the International Symposium on the Physical Problems of Colour Television, held in Paris last July. The number of copies available to non-subscribers to either journal will be limited; the size of the volume will be 400 pages with more than 450 illustrations. Advance subscriptions, should be sent to Optica Acta, 3 boulevard Pasteur, Paris 15eme, or Acta Electronica, 23 rue du Retrait, Paris xx°, before 1 January. (Price, post paid, Fr. 4,000, $10 or £4).

Winston Electronics Ltd, of Shepperton, Middlesex, have been appointed sole United Kingdom agents for a number of industrial electronic control instruments manufactured by Beckman Instruments, Munich, and Beckman Instruments Inc, Fullerton, California. The instruments have applications to gas and air compressors, coils and valves, refrigeration, the manufacture of electronic components, transformer casings, glass-metal seals. the television, valve and radar industries and manufacturers of electronic equipment, and in the engineering, gas and fluid handling and transport, aircraft and metallurgical industries.

Marconi's Wireless Telegraph Co Ltd have recently received orders from Venezuela for a complete Band 3 television station and one for a medium-frequency sound broadcasting transmitter. Other Marconi orders in South America have included a £1 000 000 telecommunications contract in Ecuador, a high-power broadcasting station in Argentina, another complete television installation in Venezuela, radar and r.t. installations for the Venezuelan and Chilean Navies and a considerable quantity of air radio equipment.

The Ministry of Supply mentions that high power radar equipment has been installed in the 45ft diameter radio telescope at the Royal Radar Establishment, Malvern to make radar observations of the Russian satellite and its rocket.

Pye Ltd have recently despatched a special Three Channel Recording Thermometer to St. Erik's Hospital, Stockholm. This is the first ordered by a foreign hospital for use in temperature investigations during hypothermia in man (low temperature surgery).

The Plessey Co Ltd announce the formation of a subsidiary company, Rosite Ltd, which will produce at its Swindon factory a wide range of cold moulded plastics for use in the electrical industry. By the terms of an agreement recently made with Rostone Corporation, Lafayette, Indiana, U.S.A. Rosite Ltd will manufacture a range of cold moulded plastics similar to that developed by the American company.

Alma Components Ltd are moving their works and offices from 165 Ossulston Street, London, N.W.1. to larger premises at 551 Holloway Road, London, N.19. Telephone Archway 0014/5.

Dr. James Reekie has been appointed Chief Engineer to Semiconductors Ltd, Ilford, Essex. For the last twelve years Dr. Reekie has been in Canada, where he was Professor of Physics at the Royal Military College of Canada. Prior to joining Semiconductors Ltd, he was Research Director in semiconductors and solid state physics for the Northern Electric Company, Montreal.

Semiconductors Ltd have announced the construction of a specially designed factory on the Cheney Manor Estate, Swindon, solely for the production of transistors. Production at the new factory is expected to start during 1958 with the manufacture of the high frequency Surface Barrier and Micro-Alloy types of transistors for military and other specialist applications currently made by the Lansdale Tube Co, U.S.A. Transistors in this range provide useful amplifier service up to 40Mc/s and can be employed for switching at rates in excess of 20Mc/s.

The merger of Elliott Brothers (London) Ltd and Associated Automation Ltd has now been accomplished. The linking of the two Companies and their subsidiaries forms the largest automation and instrumentation organization in Europe. Elliott-Automation brings under one control the design, manufacture and distribution of a very wide range of electronic, electrical, pneumatic, hydraulic and mechanical equipment, and experience in its application.

The Fourth National Symposium on Reliability and Quality Control will be held from 6-8 January, 1958, at the Statler Hotel, Washington, D.C. The Symposium will include some 45 papers by leading authorities and there will also be nine outside tours. In a period of four years, the symposium has gained acceptance as a primary forum for all concerned with reliability. Further inquiries can be made to Mr. W. H. Rombach, Chief Quality Control Engineer, Philco Corporation, 4700 Wissahickon Avenue, Philadelphia 44, Pa. U.S.A.

Aerialite Ltd have celebrated their silver jubilee this year. During the twenty-five years since its inauguration, the Aerialite Group of Companies has established a world-wide reputation in many branches of the electrical trade.

Bribond Ltd is the new name for Brighton Industries Ltd, one of the Ayling Industries Group companies, and the address is now Burgess Hill, Sussex, Telephone Burgess Hill 85611.

Erratum. On page 469 of the October issue, in the article entitled, "The Calibration of Vibration Pick-ups," by K. C. Foster, the formula embracing acceleration should read

$$a = \frac{4\pi^2 f^2 \ (A/2)}{32{\cdot}2 \text{ (ft) or } 981 \text{ (cm)}}$$

Meetings this Month

THE BRITISH COMPUTER SOCIETY

Date: 16 December. Time: 6.15 p.m.
Held at: Northampton College of Advanced Technology, London, E.C.1.
Lecture: Parallel Programming: a study of a new technique in digital computer programming.
By: S. Gill.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 18 December. Time: 6.30 p.m.
Held at: The London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: Recent Developments in Electronic Instrument Design.
By: E. Garthwaite and A. G. Wray.

North Western Section

Date: 12 December. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester, 1.
Lecture: Process Heating.
By: M. O. C. Horgan.

North Eastern Section

Date: 11 December. Time: 6 p.m.
Held at: The Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-upon-Tyne.
Lecture: Stereophonic Sound and Tape Recorders.
By: D. H. McBean.

South Wales Section

Date: 4 December. Time: 6.30 p.m.
Held at: Cardiff College of Technology and Commerce, Cathays Park, Cardiff.
Lecture: Demonstrations on Aerial Circuits.
By: H. V. Sims.

THE BRITISH SOUND RECORDING ASSOCIATION

Date: 20 December. Time: 7.15 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: Lightweight Pick-up Design.
By: Stanley Kelly.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meeting

Date: 5 December.
Lecture: Advanced Types of Power Reactors.
By: J. V. Dunworth.

Informal Meeting

Date: 9 December.
Discussion: Where Should Research End and Development Start?
Opened by: K. J. R. Wilkinson.

Radio and Telecommunication Section

Date: 11 December.
Lecture: A Flying-Spot Film Scanner for Colour Television.
By: H. E. Holman, G. C. Newton and S. F. Quinn.
Date: 16 December.
Informal Evening on: Electronics and Automation—Electronics in the Textile Industry.
Talk by: K. J. Butler.

Joint Meeting

Date: 17 December.
Held at: The Institution of Civil Engineers, Great George Street, Westminster, S.W.1.
Lecture: The Engineer and Management.
By: Sir Ewart Smith.
(Joint meeting with The Institutions of Civil and of Mechanical Engineers.

Measurement and Control Section

Date: 17 December.
Lecture: Recent Uses of Ultrasonics in Investigating the Characteristics of Materials.
By: J. Lamb.

Cambridge Radio and Telecommunication Group

Date: 3 December. Time: 8 p.m.
Held at: The Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: The Mullard Radio Astronomy Observatory.
By: M. Ryle.

North Eastern Radio and Measurements Group

Date: 2 December. Time: 6.15 p.m.
Held at: King's College, Newcastle-upon-Tyne.
Lecture: Recent Uses of Ultrasonics in Investigating the Characteristics of Materials.
By: J. Lamb.

North Midland Centre

Date: 3 December. Time: 6.30 p.m.
Held at: The Offices of the Central Electricity Authority, Yorkshire Division, 1 Whitehall Road, Leeds.
Informal Evening on: The Use of Transistors in Radio and Television.
Talks by: A. J. Biggs and E. Wolfendale.

North-Western Centre

Date: 3 December. Time: 6.15 p.m.
Held at: The Engineers' Club, 17 Albert Square, Manchester.
Lecture: Cathodic Protection.
By: L. B. Hobgen, K. A. Spencer and P. W. Heselgrave.

North Scotland Sub-Centre

Date: 12 December. Time: 7 p.m.
Held at: The Electrical Engineering Department, Queen's College, Dundee.
Lecture: Choice of Insulation and Surge Protection of Overhead Transmission Lines of 33kV and Above.
By: A. Morris Thomas and D. F. Oakeshott.
Date: 13 December. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
Repeat of above lecture.

South-East Scotland Sub-Centre

Date: 3 December. Time: 7 p.m.
Held at: The Carlton Hotel, North Bridge, Edinburgh.
The Forty-Eighth Kelvin Lecture: Infra-Red Radiation.
By: G. B. B. M. Sutherland.
Date: 17 December. (Time and place as above).
Lecture: Ship Stabilization, Automatic Controls, Computed and in Practice.
By: J. Bell.

South-West Scotland Sub-Centre

Date: 10 December. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Informal Evening on: Electronics and Automation, Some Industrial Applications.
Talk by: H. A. Thomas.

South Midland Centre

Date: 2 December. Time: 6 p.m.
Held at: The James Watt Memorial Institute, Great Charles Street, Birmingham.
Lecture: Electronic Control of Machine Tools.
By: D. T. N. Williamson.

North Staffordshire Sub-Centre

Date: 9 December. Time: 7 p.m.
Held at: The Town Hall, Hanley.
Lecture: Cathodic Protection.
By: L. B. Hogben, K. A. Spencer and P. W. Heselgrave.

Western Centre

Date: 9 December. Time: 6 p.m.
Held at: The South Wales Institute of Engineers, Cardiff.
Lecture: 138kV Submarine Power Cable interconnexion between the mainland of British Columbia and Vancouver Island.
By: T. Ingledow, R. M. Fairfield, E. L. Davey and K. S. Brazier.

South Western Sub-Centre

Date: 5 December. Time: 3 p.m.
Held at: The Electricity Showrooms, New George Street, Plymouth.
Lecture: The B.B.C. Sound Broadcasting Service on Very High Frequencies.
By: E. W. Hayes and H. Page.

West Wales (Swansea) Sub-Centre

Date: 12 December. Time: 6 p.m.
Held at: The Conference Rooms, South Wales Electricity Board Showrooms, The Kingsway, Swansea.
Lecture: Cathodic Protection.
By: L. B. Hobgen, K. A. Spencer and P. W. Heselgrave.

Maidstone District

Date: 2 December. Time: 7 p.m.
Held at: The Maidstone Technical College, Tonbridge Road, Maidstone.
Lecture: Some Medical Uses of Radio-Active Isotopes.
By: E. H. Belcher.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Date: 2 December. Time: 5 p.m.
Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Lecture: Junction Transistors and their Application to Switching Circuits.
By: R. V. Leedham.

THE TELEVISION SOCIETY

Date: 13 December. Time: 7 p.m.
Held at: The Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Dressing Television: Cabinet Design.
By: L. J. Griffin.

PUBLICATIONS RECEIVED

THE 1956-57 ANNUAL REPORT OF THE BRITISH STANDARDS INSTITUTION has recently been published. The report covers the Institution's year from 1 April, 1956, to 31 March, 1957. British Standards Institution, British Standards House, 2 Park Street, London, S.W.1. Price 7s. 6d.

THE CONSERVATION OF NATURAL RESOURCES is the title of a book published by The Institution of Civil Engineers. It contains seven lectures recently delivered at the Institution by men who have made a life study of this particular subject, and who view with alarm the rapid expenditure of resources such as metals, land, and fresh water, which increasing industrialization is bringing about. The Institution of Civil Engineers, Great George Street, Westminster, London, S.W.1. Price 7s. 6d. to members and 17s. 6d. to the public.

TELEVISION TRADE CATALOGUE, CONSUMER CATALOGUE, COMPREHENSIVE TRADE CATALOGUE and an AERIAL WALL CHART giving a comprehensive list of all television aerials, designed for easy reference, are recent publications of Aerialite Ltd, Castle Works, Stalybridge, Cheshire.

FIRST TRANSATLANTIC TELEPHONE CABLE is the subject of a booklet issued by Submarine Cables Ltd, of Mercury House, Theobald's Road, London, W.C.1. This is a company owned jointly by Siemens Edison Swan Ltd and The Telegraph Construction and Maintenance Co. Ltd.

A SECOND SELECTION OF EQUIPMENT FOR THE LABORATORY by Kendall and Mousley, of 18 Melville Road, Edgbaston, Birmingham, 16, has been published at a price of 2s. 6d. The book is also available in Australia from Kendall and Mousley Ltd, 59 Scottish Avenue, South Road Park, South Australia.

BRITISH NUCLEONIC INSTRUMENTS 1957 is a book which has recently been published by the Scientific Instrument Manufacturers' Association of Great Britain Ltd, SIMA House, 20 Queen Anne Street, London, W.1. It is the fourth edition and describes the great majority of the instruments which are available from British firms.

FERRITE MATERIALS is the title of a catalogue issued by Siemens & Halske, of Munich, for whom R. H. Cole (Overseas) Ltd, of 2 Caxton Street, London, S.W.1, are agents for these materials in this country.

THE PRACTICAL ELECTRICIAN'S POCKET BOOK 1958, edited by Roy C. Norris, has recently been published. This is the sixtieth edition. Electrical and Radio Trading, 189 High Holborn, London, W.C.1. Price 5s.

ASTRONIC is the title of a catalogue dealing with the small power amplifiers, transformers and chokes manufactured by Associated Electronic Engineers Ltd, Dalston Gardens, Stanmore Middlesex.

NEWNES ELECTRICAL POCKET BOOK, edited by E. Molloy, fourteenth edition. This edition has been thoroughly revised so that the information given is in accordance with the requirements of the thirteenth edition of the I.E.E. Regulations for the Electrical Equipment of Buildings. In addition, there is a new Section, summarizing the regulations and giving useful tables on Cable Sizes, Diversity Factor, Capacity of Conduits and Fuse Wire Values. George Newnes Ltd, Tower House, Southampton Street, Strand, London, W.C.2. Price 10s. 6d.

UNIVERSAL AND SPECIAL PURPOSE ELECTRONIC DIGITAL COMPUTERS is a leaflet which describes the DEUCE universal digital computer and also special purpose computers. The English Electric Co. Ltd, Queens House, Kingsway, London, W.C.2.