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Commentary

THE British Institution of Radio Engineers recently held a convention at King's College, Cambridge, and had as a title "Electronics in Automation". The convention was attended by about 500 persons, including delegates from some eighteen overseas countries; one of the most interesting contingents being a party of six from Russia. During the course of the convention forty or so papers were read and discussed and most of the more important applications of electronic engineering to increased or better production were covered in some form or other.

The greatest value of a meeting of this type is not, however, the reading of the papers, for, after all, these will eventually be published and can be read at leisure. The most important part of a convention is the discussions which take place, both the formal ones in the lecture theatre and the informal ones which take place in the delegates' spare time. This is not to say, of course, that the papers are of no importance, but they are, so to speak, a nucleus around which the convention can revolve and not an end in themselves. The opportunities for the meeting and interchange of ideas between engineers from different parts of the country, and from different countries, which a meeting of this type provides are of the utmost use and value and, if the delegate makes the most of these opportunities, the time spent away from the laboratory can prove not only well worthwhile but a good investment.

Some of the most thought-provoking statements were made during the final summing-up session of the convention and, of these, two are worthy of particular note. The first was made by the Chairman, Dr. A. D. Booth, who, talking about computers, stated that a fundamental new philosophy is needed towards the design of business machines and that the philosophy of the designers of five years or so ago, who were concerned with machines intended primarily for scientific computations, is no longer valid. Certainly it is true that, although the various techniques may be used with more finesse and elegance, there have been few fundamental advances in the last five, or even ten, years. In the case of memory devices, for instance, the magnetic core dates back to about 1947, a magnetic drum was used as long ago as 1900 for recording telegraph signals, while the Williams cathode-ray tube store originated about 1943. Thinking on these lines it is, perhaps, a sobering thought to realize that the transistor has now been with us for some ten years. There is a variety of computers now being produced specifically for business use, but the number in use is not large and it may well be that before these machines are accepted as an intrinsic part of a business organization a completely and fundamentally new approach to their design will be called for.

The second statement was in fact no more than a passing remark which was made by Mr. H. J. Finden, who said words to the effect that, although a lot of ideas start in this country, we are often slow off the mark in developing and exploiting them. This, unfortunately, is very true and has been true for a very long time. It is a point which was made quite forcibly by the Marquis of Salisbury when he was opening the new Ekco laboratories a month or two ago. The Marquis quoted an example completely divorced from the electronic industry but one of somewhat personal significance; that of aniline dyes. These dyes were, in fact, discovered, or invented, by the Marquis's grandfather but they are now almost universally regarded as of German origin. It is not difficult for even the young engineer to look around and see examples of similar happenings in his own sphere. Why this should occur so frequently is by no means clear. Whether it is that British scientists are less commercially minded than those of other nations or whether those responsible for financing new projects have too conservative an outlook and not enough confidence in the scientist. It is probably due, in part, to the attitude which is engendered by the type of academic institution which has developed over the centuries in this country and wherein the pursuit of absolute knowledge tends to blind the eyes of the investigators to the practicalities of everyday life. Whatever the cause it is certain that, although 'research for research sake' may be a fine ideal, it does not pay the kind of dividends that a country needs for its existence in the modern world. Those engaged on research at any level should be encouraged to envisage and, most important, bring to notice, the practical applications of all they do: while those responsible for the commercial development of the fruits of research must play their part with courage and foresight.

The Design of the Pegasus Computer

(Part 1)

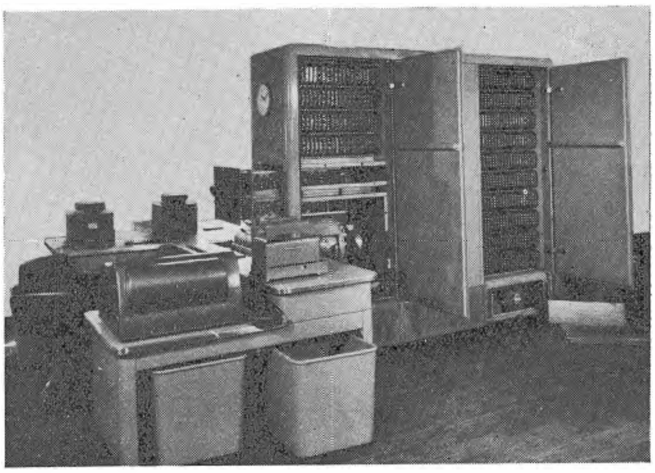
By T. G. H. Braunholtz*

It is three and a half years since the structure and Order Code of Pegasus were decided on. This article records some of the background of opinion from which the design was thought out. Only topics which can be discussed simply have been included, and the article is intended to be comprehensible to readers with only a slight knowledge of computers.

In Part 2 of this article the more detailed engineering aspects will be dealt with, with particular emphasis on the implications of building the machine from standard plug-in packages.

THE Ferranti Pegasus Computer is a medium-sized general-purpose digital computer, designed by Ferranti Ltd in collaboration with the National Research and Development Corporation. To give a general picture of the scale of the machine: the store consists of a drum of 5 000 words and an immediate access store of 55 words, the multiplication time is 2msec, the basic operation time is 0·3msec, and the digit repetition rate is 330kc/s. It is perhaps worth remarking that, as a guide to the overall speed of a computer, none of the figures given is any use by itself. Speed can only be judged by considering the whole structure of a computer.

It is clear by now that a second stage in the development of digital computers is underway, marked by a serious endeavour to make them less difficult to use. Since the Pegasus goes some way in this direction and contains a number of novel features, an account of the ideas behind the design may be of interest.



A typical arrangement of the Pegasus computer

This article is limited to the less technical and detailed aspects of computer design, and does not attempt to discuss points which can be of interest to programmer's only.

After a few remarks on the cost and reliability of computers, there follows a discussion of the principal subject of this article, the design of a digital computer from the programmer's point of view. This is divided into two sections, of which the first is more general, and the second is concerned with the order code.

Minimization of Cost

A considerable part of the cost of the more expensive computers is due to such features as very fast multipliers and shifts, and very large immediate-access stores. These have been dispensed with in Pegasus, but the design is such as to minimize the loss in convenience and speed arising from their exclusion.

Reliability, of course, takes precedence over cost. However, the cause of unreliability lies generally in a few weaknesses in the machine design, and though a great deal may be spent in investigating and eliminating these weaknesses, their correction is not generally very expensive in equipment. Thus, with well developed techniques, cost and reliability are not in serious conflict with one another.

Ease of use in a computer is, too, as much a matter of careful analysis of the problems of programming in relation to the design of the computer, as of the size and cost of the machine.

Reliability

An unreliable machine involves the user in much more trouble than simple doubt as to the accuracy of the solutions obtained. A programmer using an unreliable machine wishes to ensure checked results even when occasional errors in computation are occurring. To do this he divides his programme into sections each of which is checked separately and can be repeated if the check fails. His programme consequently becomes far longer and more difficult to code; its speed drops and its complexity rises. A programmer can easily involve himself in more thought and effort in executing his safety measures than in the whole of the rest of his problem.

With a really reliable machine, checking can be simple. The question naturally asked by electronic engineers when computers were first built was how reliability was to be achieved. That problem has long since been largely solved. It is more pertinent today to ask why reliability is sometimes not achieved, and two principal reasons can be found.

Firstly, some of the computers which have been in use for several years have been prototype models incorporating new electronic techniques for experimental purposes. Naturally these were not often completely reliable in their first embodiment, and it may have been impossible to carry out the needed modifications within the framework of the existing computer.

In view of this, it has been the policy in designing Pegasus to use only existing and well-tried engineering techniques.

Secondly, there is a great temptation to solve difficult design problems by lowering the safety margins. But in computers the consequences of errors are so serious that wherever there is doubt as to the performance of a component, the most pessimistic view should be taken. It is tempting to reduce safety margins because engineering problems would be greatly eased by doing so, but the temptation has to be resisted.

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Programmer's Analysis of Machine Design

In this section design from the programmer's point of view is discussed in some detail. The broad aims in design were to achieve

(a) Ease of programming.

(b) Ample capacity for handling general engineering and scientific problems.

(c) High speed of computation.

MAIN STORE CAPACITY

Opinions as to a desirable size of main store are bound to be at variance with one another, but it seems wise at present to err on the side of too much storage rather than too little. While it is true that almost any amount of storage seems to be profitably used if it is available, it is our experience that 1 000 words would be inconveniently small, that 2 000 words might possibly be sufficient, and that 4 000 words should be ample, especially if magnetic tape is used in addition.

The main store of Pegasus is a magnetic drum with a capacity of 5 120 words of 40 digits each. The drum consists of 40 tracks containing 128 words each. Since the speed of rotation is 3 750 rev/min the maximum access-time is 16msec.

OPTIMUM PROGRAMMING

It is possible in machines with a magnetic drum main store, or any other form of storage with a long access-time, to arrange that orders are obeyed directly from that store. If, however, the drum revolution time is long in comparison with the basic order time, the orders should be placed so that it is only rarely that a drum revolution is wasted through waiting for an order or a number. The accomplishment of this is known as optimum programming. An alternative is to have sufficient immediate-access storage (or, at any rate, quick access storage) to be able to carry out all operations in the fast store. The size of fast store required, both for ease of coding, and for speed of computation, can be made surprisingly small if access to the main store is made easy.

Optimum programming, on the other hand, gives rise to many difficulties. It has been stated half seriously that for a reasonable expenditure of effort by the programmer on optimum programming, the average time spent in obeying orders is proportional to the square root of the revolution time of the drum (the unit of time being the basic operation time of the machine). With optimum programming, the achievement of high speeds requires a great deal of thought and ingenuity, which, though presenting an attractive challenge to the programmer, does not lead to good use of the computer. A highly developed use of sub-routines is exceedingly difficult with optimum programming, and moreover it is found that when more than about half the available storage is occupied by a programme and its data, the difficulties in optimum programming become very great. These last two remarks apply particularly to forms of storage with a maximum access-time greater than 32 word-times. With 32 word-times or less, efficient sub-routines can be constructed, and an efficient use of the storage can and has been achieved.

For these reasons optimum programming was rejected as unsatisfactory for Pegasus.

THE COMPUTING STORE

By providing an immediate-access computing store in which all operations are carried out, it has been possible to avoid optimum programming in Pegasus. This computing store contains 55 words of 39 digits each. There is immediate access to each word.

Each word contains either one number or two orders. All orders are obeyed from the computing store and all arithmetic operations are carried out on numbers held in the computing store.

Transfer between the main store and the computing store is either by eight-word blocks or a single word at a time.

The reader will probably feel surprised that 55 words should be sufficient for general programming purposes. He will convince himself most easily by trying programming for himself, but some examination of the problem may be helpful.

USE OF COMPUTING STORE

For general usefulness, a more important factor than the absolute size of the computing store is the number of blocks into which it is divided for purposes of transfer to and from the main store. We have found that six blocks, each of eight words, in addition to the seven accumulators, is ample. Transfer between the computing store and main store is normally in eight-word blocks (single-word transfers being used for storage of rarely used counters when counter space is short). Generally, three blocks will be used for data and results, one for counters and other variables, and two for programme. Often this allocation is not kept to, and in simple programmes several blocks may well be left unused.

The change-over from one block of programme to the next, which might be expected to be a serious difficulty, may be accomplished very simply. Two orders are picked up at once (i.e. in one word), so that if the first is a block-transfer and the second a jump order, then the sequence of operations will be as follows: the new block of programme will be brought down over the old (which is therefore obliterated) by the first order; then the second order, which is a jump, sends control to any position in the new block that is desired. This allows each block to be coded quite independently of the others.

SPEED AND INNER LOOPS

Generally 80 per cent or more of the time in a computation is spent in one or two inner loops. Consequently the speed of the inner loops determines the speed of the programme. One would like these inner loops to be held entirely in the fast store, and be run through perhaps eight times (a likely number since each block of data contains eight words), until further transfers of data become necessary. In fact this can almost always be done, and in this case one may expect half to one-third of the time to be spent on block transfers. The time spent in the rest of the programme will be small, and will probably consist mostly of block transfers, and we conclude that the high operating speed of the machine will at any rate not be smothered by block-transfer time.

THE SIZE OF THE COMPUTING STORE

The effect of variations in the size of the computing store may now be considered.

It is an interesting fact that there is little advantage to be gained by increasing the size of the fast store if the size of blocks is kept constant. However, if the size of blocks were increased to, say, 12 or 16 words and there were still six blocks a general improvement in the machine's performance would result, but the size of the computing store would be nearly doubled, which would be very expensive.

If the size of blocks were decreased to four words, not only would the machine be considerably slowed down, but programming would become more difficult because of the small number of orders held in each block.

It might be that the machine would be faster and easier to use if the number of blocks were increased from 6 to 8. However, to have more than 8 would only add to programming difficulties without a compensating gain, until the computing store were made large enough to carry out a computation almost without reference to the main store. This point is naturally enough rather ill defined, but is perhaps in the region of 500 to 1 000 words.

Problems of Two-Level Storage

Two levels of storage are necessary if one is to get both a fast machine and an ample store at a reasonable price. The double store does complicate programming but the complication has been reduced to a minimum by special facilities in the order-code.

The Order-Code

Given a small computing store of the sort described, it is clearly very important that the order-code should allow concise coding. Quite on its own merits, however, it is felt that a very full, carefully organized order-code is essential to a computer. The coding and testing of a programme is a very lengthy process, and great increases in effectiveness may be gained by simplification of coding.

The Problem of Coding

In applying a computer to a problem, the first stage is to decide on the mathematical method to be used. The next stage is to prepare a 'flow-diagram' (describing the operations to be carried out, broken down into small units). The third stage is the translation of the flow diagram to coded sheets. The following discussion is limited to the problems that arise during that stage. In fact, however, each stage affects the other stages, and, since alterations are always made as the structure of the problem broadens out and clarifies, each stage reacts on the previous stages, so that the problem is so interconnected that it cannot accurately be considered as broken into separate stages.

There is little agreement as to what constitutes a useful coding facility, and what a coding difficulty. It seems that programmers tend not to be aware of the amount of effort they are diverting from their main problems to the coding of awkward points. For, when coding, the full powers of a programmer are called forth to weigh, compare, co-ordinate and to mould his programme into an elegant and unified whole.

If this is not done, then alterations, which will certainly be made as the work develops, will each add to the illogicality of a rapidly complicating programme. The final results will be errors in programming, and unreasonable interrelations and particularities in the programme which make it slow, difficult to use, and difficult to adapt.

An effect of coding difficulties more serious than the obvious waste of time involved is that the programmer's attention is diverted from the more fundamental features of his problem to tricky points of coding. As a result the fundamental features may not get the attention they deserve.

Since the greater part of a programme is concerned with organization, particular attention should be paid to easing its coding.

Coding Difficulties

A coding difficulty is defined as a point in coding which takes much of the programmer's attention, and which might reasonably be avoided by an alteration or addition to the code. An experienced programmer will naturally develop his ability to overcome coding difficulties with little effort. It is therefore particularly important to avoid coding difficulties in a machine to be used much by part time or inexperienced programmers.

Coding difficulties arise from (among other things)

(a) *Exceptions;* that is, such details of the actions of orders as have no reason obvious to the programmer for their existence, which are not automatically remembered, and which are not common to a whole series of orders.

(b) *Confusion,* as when two orders are very similar and apt to be confused.

(c) *Lack of independence,* arising when one order affects others in ways that are not naturally in the programmer's mind.

(d) *Complexity,* as in subtle codes, the possibilities of which delight many a mathematician, and in which remarkably concise sequences are often possible by sufficient thought.

(e) *Ambiguity,* as when an operation can be carried out by two sequences of orders about as good as one another: it is only human to waste effort in selecting the best. To do so is not as over-meticulous as might appear, since the programmer will feel that the decision reached will have further application whenever in the future a similar situation arises. The trouble is that the number of such situations is very great in certain codes.

(f) *Awkwardness,* for example, when floating-point operations are lengthy to code. (Many troubles of awkwardness can be avoided by a well organized system of sub-routines.)

A few particular conditions which it is felt a code should satisfy are listed below:

A code should be easy to remember, both to aid the actual coding, and to facilitate reading and checking the coded sheets; so that it may be easy to remember, the orders should be arranged according to some simple rules. Particularly, if there are many jump orders (as there should be) these must be arranged logically. Again, the code should be 'consistent'—there should not be gaps in the series of arithmetical or jump instructions.

A code should avoid unsigned numbers; the gain of one digit is small compared with the confusion created.

A code should allow simple working with both integers and fractions; allow simple programming of such operations as floating-point and double length arithmetic, and include many jumps in pairs with contrary sense. Jumps and switches in organization can then be arranged without much doubt as to the best coding.

The orders and their details should be as simple as possible.

It is perhaps not often realized how little extra equipment a full order-code needs beyond the requirements of a bare minimum code. The total equipment in a machine is certainly not increased by as much as 10 per cent.

Provision of Multiple Accumulators

This is one of the unconventional features of the design of Pegasus; it is based on a detailed examination of various possible order-codes.

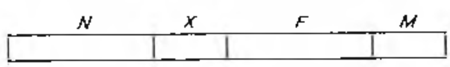
An obvious starting point is the three-address code, with orders of the type $A+B \rightarrow C$. An examination of the use of this code shows that very often two of the addresses are the same, i.e. the order takes the two-address form $A+B \rightarrow A$. A further examination shows that in a high proportion of these cases the address A which occurs both as a source and destination is confined to a very few numbers. This leads to the suggestion that a restricted two-address code of the type $X+N \rightarrow X$, where X ranges over only a few addresses while N can be anywhere in the computing store, may be a very economical type of code.

An alternative starting point is the common single-address code. It is found generally that one-third or more of the orders are used simply to shunt numbers to or from or via the accumulator. This is because in coding one often carries out a few operations building up one number, and then transfers attention to other numbers, meanwhile holding the first number in store for further use. Hence the large amount of shunting. It seems reasonable, therefore, to provide several accumulators, in which the partially processed numbers may be left. One is thus led by simple arguments from two opposite extremes to the type of order code adopted in Pegasus.

Modifying registers are considered indispensable, and by giving the accumulators the additional function of modification, facilities for arithmetic operation on the modifying registers are automatically provided.

The orders in the multiple-accumulator code are moreover very concise, so that it is possible to get two orders comfortably into a word of reasonable length.

The arrangement



of the instruction is the most convenient from the engineering point of view. (Usually N is the register address, X the accumulator address, F the function and M the modifier address).

Word and Order Length

For a general-purpose computer the most satisfactory number of digits in a word lies between 35 and 40. 35 digits seems to be a minimum for comfort, any excess over 40 superfluous.

In order to economize in storage space it is very desirable to have two orders per word, and the form of order in Pegasus has been chosen partly with this in mind. Each order contains 19 digits, and the two orders together with the 'stop-go' digit give a word length of 39 digits. (The 'stop-go' digit, when it is '0', causes the machines to stop before obeying the order-pair in the same word.)

Parity Checks

A parity check digit is a digit associated with a number or character, and chosen so as to make the total number of ones in the number odd (or even). If the total number of ones is made odd it is called an odd parity check, and if the total number of ones is made even it is called an even parity check. The parity digit of a number, once formed, is carried about with it and is used for checking that the number is stored and transferred correctly.

The parity digit check is not a complete test for the correctness of a number, because a change of an even number of digits would not be detected. But used to its best advantage, it will detect perhaps 99 out of 100, or even 999 out of 1 000—we do not know what proportion of the storage and transmission faults which occur, before they have caused computational errors in data output from the machine. Storage and transmission faults might constitute 10 per cent or 20 per cent of the total number of faults which occur in the computer, and so by eliminating the storage from the part of the machine to be investigated when the computer is known to have a fault, and by saving the time which would be spent in tracing storage errors, an important reduction in the time spent hunting faults could result.

In Pegasus, all words on the drum and nearly all nickel lines of the computing store have a parity check digit, and also a parity digit can be formed during input or output of a character, and it is up to the programmer to make proper use of it.

Whenever a number is read from the drum its parity is checked. A failure stops the machine. The parity check is odd, and the number of bits in a word, including the parity bit, is even, so that the three most important types of faults will be detected:

(a) A single digit error, due perhaps to bad recording on the drum.

(b) The number and parity digit all zero, probably due to an electrical or electronic failure.

(c) The number and parity digit all ones, probably due to an electrical or electronic failure.

Notice that had the parity check been even, then all zeros would not be detected; and also, if the parity check is odd, the total number of digits including the parity digit must be even if all ones are to be detected.

The parity digit on the drum is particularly important because the only alternative check on storage is regular testing by programme, which is lengthy, and not easily made comprehensive.

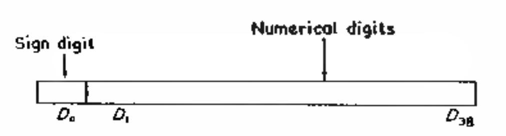
Some Details of the Order-Code

Binary-Point Convention

Consider, from first principles, the question of how the digits stored in the computer are to be interpreted. It is in fact up to the programmer to decide, and the computer only comes in because if the digits are interpreted in a particular way the orders of the machine carry out simple arithmetic operations. But several other interpretations are regularly used. One, for instance, is as orders for the use of the machine's operational circuits. Another interpretation is as orders in an 'interpretive scheme'—in which the orders are decoded and obeyed by sub-routines constructed by the programmer. A third way is as numbers. In this case the numerical value to be attached to the individual digits, that is, the position of the 'binary-point', has to be chosen. It is clear on reflection that the only operation in which the position of the binary point affects the interpretation of the action of the machine are multiplication and division. However in describing the order-code some convention is needed, and the following has been adopted as being the most natural, and the most frequent in practical use:

Numbers (in digits D_0 to D_{38} of a word) are considered to lie in the range $-1 \leqslant x < +1$ except where otherwise stated.

The digit positions of a word are numbered as below:



On this convention, therefore, the D_1 position represents 2^{-1}.

Fractions and Integers

An alternative is to interpret numbers as integers. In this case the D_{38} position represents unity. In Pegasus it is as easy to work with integers as with fractions, since multiplication is arranged so that for both, the binary point remains in its natural position; division is independent of the position of the binary point provided the quotient is interpreted as a fraction.

Overflow

The value of this overflow facility is that it gives a simple test, which can be applied at the end of any stage of a calculation, as to whether any number has run out of scale during that stage.

When an attempt is made to transfer to the main store a number which has overflowed, the machine will stop. This facility should prove useful in locating errors of mathematical reasoning behind the programme, since these are otherwise extremely hard to find.

Double-Length Numbers

Double-length numbers occupy two words of storage. The more-significant word is interpreted normally. The less-significant word is given a zero sign-digit D_0 so that it represents a positive number, as it should. The numerical digits then carry on in decreasing significance from where the more-significant word left off; that is, when the double-length number is interpreted as a fraction, D_1 of the less-significant word represents 2^{-39} instead of 2^{-1}.

Writing p, q for the standard interpretation of registers P, Q (accumulators 6, 7) and writing the double-length fraction formed from them $(p\ q)$ with the q the less significant, one has

$$(p\ q) = p + 2^{-38}\, q.$$

This is the standard interpretation of double-length numbers. The alternative interpretation as an integer is similar:

$$(p\ q)_{\text{integer}} = 2^{38}\, p_{\text{integer}} + q_{\text{integer}}$$

Pegasus permits simple working with double-length numbers principally through the following facilities:

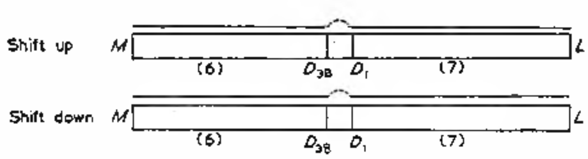
(a) Accumulating double-length products.

(b) Double-length shift and normalize in registers 6 and 7.

(c) The 'justify' order.

(a) It is often required to add together a series of products, as in matrix algebra. Although the result may only be required to single-length accuracy, nevertheless, in order to prevent loss of accuracy due to rounding errors, it is often desired to accumulate the double-length products.

(b) Shift and normalize applied to accumulator 7 is skipped in shifting, as shown:



(c) In double-length addition, for example, when the two less-significant halves are operated on, overflow may occur into D_0 of accumulator 7. This '1' should really be added in the least significant digit position of accumulator 6.

The 'Justify' order does this, and restores D_0 of accumulator 7 to zero.

The total effect of all these facilities is to make the handling of double-length numbers very simple; in particular, the problem of carry from the less-significant to the more-significant half of double-length numbers is solved with very little equipment and without introducing new programming difficulties.

Counting

The methods of modifying and counting on Pegasus are two of the novel features of the machine. The methods adopted have been developed especially to simplify the organizing sections of programmes, and will be found extremely convenient to use.

The counting needed to decide when to write-up or bring down blocks of the main store, and when to end cycling around the loops of a programme, makes a special use of accumulators in the following way. The accumulator contains two numbers, one occupying D_1 to D_{13}, the other D_{14} to D_{38}. Both numbers are thought of as integers.

The number in D_1 to D_{13} is called the modifier. It is used to keep note of where the machine has got to in the main store when running through data and to decide when to transfer information between the computing store and main store.

The number in D_{14} to D_{38} is called the counter. It is used to count the number of times the machine has cycled round a loop, and so to determine when to stop cycling.

There are two orders which make it simple to carry out these operations. These are described later.

Counter-Orders

These orders are used particularly in setting and altering counters.

Each ordinary arithmetic order has an analogous counter order; both sets of orders can thereby be more easily remembered.

In a counter-order, the integer formed by the N-digit of the order is one of the operands; the counter in the accumulator specified is the other operand, and the result of the operation is placed in the same accumulator. Thus if N contains the integer 13, then 13 can be added, subtracted etc. from the counter specified in the X-digits.

Modification

If it were not for the fact that in long mathematical problems the same series of operations have to be repeated very many times, then digital computers would be of little value for such work, a fact which has often been pointed out. But this repetition does occur, with slight regular alterations carried out by the sequences of orders on themselves each time round the cycle. In most early computers the alterations—which generally consist of adding or subtracting small integers from the addresses of orders—were carried out on the orders themselves, but in the Manchester University computer the idea of modification was introduced. The contents of certain registers, the B-lines, could be added to an order before it was obeyed without altering the orders in their store position, so that by altering the B-line the appropriate modification of the sequence of orders was carried out.

This system has three major advantages.

First, the alteration of a B-line is very much simpler than the alteration of an order, particularly since orders are often over-written on the Manchester machine (both it and Pegasus have two-level storage).

Second, one modifier can often be used to modify several orders in a sequence.

Third, the counter can also be used to test for the end of the loop.

In Pegasus modification has been carried a stage further; different orders are modified differently according to their needs. In particular, the difficulties of organization associated with two-level storage have been greatly reduced by special modification arrangements.

There are four types of modification:

(a) *With arithmetical orders*, in which one is interested in modifying an address within a block of eight, modification adds to N the part of the modifier less than 8.

(b) *With shift, normalize, and counter orders*, modification adds to N the whole modifier (this is not often used).

(c) *With block transfers*, one is not interested in the part of the counter less than eight, since this refers to a position within a block. It is the rest of the counter which is important, so the modification adds the block number part of the modifier to N.

(d) *With single-word transfers*, in which the whole counter may be used to specify the word, and in which the N and X digits of the order are used to specify the word, the whole modifier is added to the N and X digits.

'UNIT MODIFY'

The unit modify order adds one to the modifier in X and jumps to N if the three least-significant digits of the modifier are not all zero. This order makes it simple to carry out a routine process on consecutive numbers running through several blocks of the main store, since it moves on the modifier by unity each time, and also determines when transfer of data to or from the main store is required.

'UNIT COUNT'

The unit count order subtracts one from the counter in X, and jumps to N if the counter is not zero. It is used to count the number of times the machine carries out a routine process. An illustration of the use of these two counting orders is given:

Initial Setting etc.

→ Process one word (use modification (a))

— Unit modify and jump if the 3 least significant digits of the modifier are not all zero

if≠0 ↓ if=0

Read next block (use modification (c))

→ Unit count and jump if the counter is not zero.

if≠0 ↓ if=0

End of Process

Note. The jump after unit modify will occur eight times between runs through the block read order. Hence the modifier will have eight added to it between one order and the next, so that the block number in the modified transfer-order will increase by one each time round.

Acknowledgments

The author would like to acknowledge his debt to Mr. C. Strachey, to whom the form taken by the Pegasus computer is largely due.

Acknowledgment is made to Ferranti Ltd for permission to publish this article.

(To be continued)

A Photocell Counter for Fertilizer Loading

At the Barking factory of Fisons Ltd, a simple form of electronic photocell counting is employed in conjunction with their multi-wall paper sack filling machine to give the aggregate count of sacks per shift, the load per lorry and the number of burst sacks.

The 1cwt sacks are of the valve closure type and are filled at the rate of 600 per hour with a total tonnage of 30 per hour. There are no checkers or tallymen on the lorries, everything being controlled from the filling machine on the first floor where there are two men only. One to operate the two head filling machine and one to keep him supplied with sacks, take off 'bursts' and switch the chute guide as each lorry load is completed: there being two lorry 'stations' at the loading bank.

The filling machine operator places a sack on the head, operates the hopper valve and allows the sack to be automatically filled with exactly 1cwt. There is a vibrator foot on which the sack rests which causes the fertilizer to be shaken down during the filling cycle, thus allowing an even filling of the sack which is then allowed to drop on to a conveyor while the operator is working on the next sack.

The conveyor then takes the sack through the 'counting tunnel' in which are situated the photocell light heads. The sack passing through breaks a light beam causing the 'count' to be registered.

The sack now passes to the chute which takes it to the ground floor and thus to the lorry at the loading bank. Display counters are remotely fitted and the first display counter registers the lorry load total and completion of this total is the signal for the second operator to switch the chute guide so that the second waiting lorry can commence loading, the second display counter registering the aggregate for the shift.

Inevitably there are a certain number of burst sacks and the total number of 'bursts' at any time can be quickly ascertained by use of this equipment.

The electronic counting equipment was supplied by the Electronic Machine Co. Ltd.

An Electronic Calculator for Harwell

The first Type 555 Electronic Calculator, made by The British Tabulating Machine Company Limited, is now in use by the Computing Group of the United Kingdom Atomic Energy Authority at Harwell.

The machine will be employed almost exclusively for data processing and for computations involving use of the 'Monte Carlo' technique. The first concerns the interpretation of measurements made with experimental facilities attached to the research reactors (BEPO, DIDO, etc.) and to the various high energy particle accelerators.

Nuclear physicists at Harwell are undertaking very large programmes of measurements of fundamental properties of matter, essential to the proper development of reactor technology, as well as to the understanding of nuclear physics.

The whole of the process is being mechanized by means of Hollerith power punches, directed and controlled by the apparatus concerned; by which readings are automatically recorded in punched cards. Calculations are made by the calculator, at the rate of 6 000 card passages per hour, and their results are recorded in punched hole form. These can subsequently be printed out by tabulator machine, or produced in graph form by a Dobbie McInnes card-controlled plotter.

The Monte Carlo mathematical-statistical method for studying physical phenomena in which random processes occur is being employed with the aid of the electronic calculator to trace the life history of neutrons, entirely by calculation. The first problem posed concerned the leakage of neutrons down channels in a reactor, as, for example, those which carry cooling ducts. The method involves tracing the life histories of several thousands of neutrons—a sample survey—and using this information to predict the behaviour of the very much larger number of neutrons in the actual reactor. This type of problem, intractable by conventional mathematical methods, because of the huge amount of computation necessary, now becomes capable of solution with the assistance of high-speed electronic calculating machines.

A CATHODE-RAY TUBE OSCILLOSCOPE FOR USE IN MILLIMICROSECOND PULSE TECHNIQUES

By J. W. Armitage*, A.M.I.E.E., G. Gaskin* and K. Phillips*, M.Sc., A.M.I.E.E.

This article describes the construction and performance of an oscilloscope which can be used for observing high speed transient phenomena in the millimicrosecond region. The time-base employs a thyratron trigger and a hard valve sweep circuit with a continuously variable sweep speed from 3μsec down to 15mμsec with a scan of approximately 2kV. A 908BCC sealed-off cathode-ray tube is used. A pulsed calibrating oscillator is also described which operates up to 350Mc/s.

IN the fields of gas discharge and nuclear physics research, techniques have been developed which involve the production and observation of very short electrical pulses with durations of less than a microsecond and as short as a few millimicroseconds. In the past, several workers have described continuously evacuated cold-cathode tubes operating with high accelerating voltages of the order of 50kV and in the majority of cases the electron beam was allowed to impinge directly on to the recording photographic emulsion. However, with the development of improved sealed-off cathode-ray tubes these continuously evacuated oscilloscopes with their cumbersome pumping equipment are now almost out of date; except in the case of the micro-oscilloscope[1] which is used for ultra high speed oscillography. One obvious advantage of a cold-cathode tube is that its life is almost unlimited and furthermore, in the investigation of high voltage transient phenomena it is often possible to apply the voltage directly to the deflector plates. This eliminates the potential divider which is often the cause of considerable transient waveform distortion. The sealed-off tubes are invariably more compact and consequently more portable. These are only some of the relative merits of the sealed-off and continuously evacuated instruments which have been discussed at great length by other workers. When designing an ultra high speed oscillograph these relative merits must be considered. A further important consideration is the ability of the cathode-ray tube to record high speed transients. The faithful reproduction of a high frequency waveform on the screen of an oscilloscope is limited by the cut-off frequency of the signal deflector system. This is fixed by the transit time of the electrons through the deflexion system and also by the distortion caused by the impedance mismatch between the deflector system and the signal line. The cut-off frequency can be increased by making the conventional deflexion plates into a travelling wave system or into a twin wire system which can be matched to the impedance of the signal lines. With the former, deflexion sensitivities comparable with the normal plate system can be obtained while with the latter system the sensitivity is greatly reduced but, as in the design of most instruments, a compromise has to be made. This low deflexion sensitivity is not very important when high voltage pulses are used, since it only means that a lower ratio potential divider has to be used. However, for lower voltages (say for use with a photomultiplier) a lower sensitivity oscilloscope usually means that amplifiers having a high gain, together with large output voltages have to be employed and in the millimicrosecond range this is not always very convenient. Since this cathode-ray oscilloscope was constructed as a general purpose laboratory instrument an attempt was made to make it as versatile as possible. The 908BCC cathode-ray tube was selected for this purpose, since it has a deflexion sensitivity for the signal plates of about 90V/cm. This

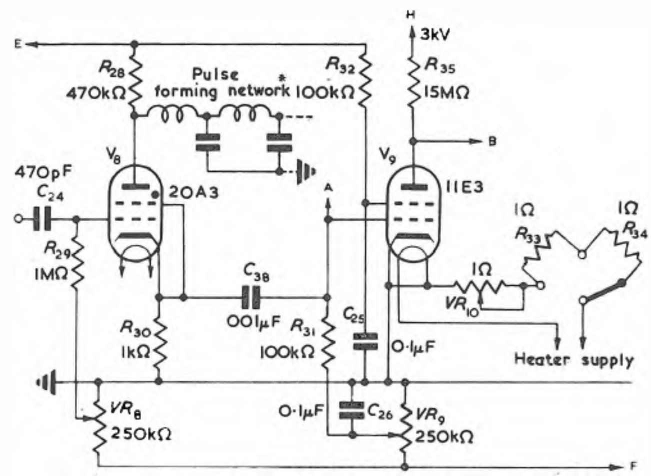


Fig. 1. Time-base circuit
* *Pulse forming network consisting of 12 sections of inductance 0·125mH per section and capacitors of 500pF*

tube has side arm connexions to the deflecting plates limiting the effect of crosstalk, and reducing the capacitance between the connexions and also the loop inductance with the result that the natural frequency of the circuit is in the region of 300Mc/s. The transit time of the electrons through the signal plates is of the order of $7 \times 10^{-4}\mu$sec and the attenuation at 500Mc/s is only about 10 per cent[2].

Description of the Circuits

The Time-base Circuit

Several types of time-base circuits were investigated. In the past hydrogen thyratrons have been used, but these were abandoned owing to their long firing delay (0·25μsec). Delays in the time-base circuit invariably involve delaying the signal pulse with a length of coaxial cable. The delay cable attenuates the higher frequency components of the pulse with a consequent distortion of its shape, and hence it is desirable to reduce the length of cable to a minimum. Attempts were made to reduce this delay by triggering the hydrogen thyratron with rapidly rising trigger pulses applied to its grid, but it was found impossible to reduce

* *Metropolitan Vickers Electrical Co. Ltd.*

this delay to much less than 0·1μsec. In an attempt to avoid these inherent delays in thyratrons several hard valve time-bases were tested[4]. The circuit finally used (see Fig. 1) was similar to the one originally suggested by Smith[5] for use with a micro-oscillograph.

The time sweep valve, V_9, an 11E3 pulse modulator

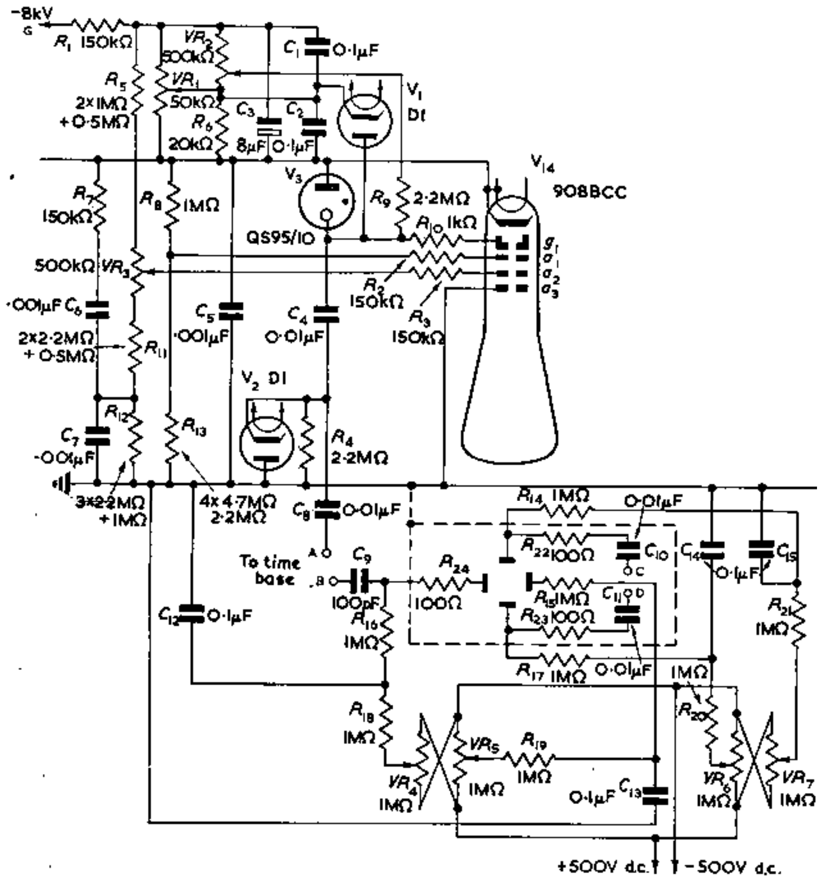


Fig. 2. Beam modulating and c.r.t. protection circuit

tetrode, is normally cut off, then rapidly driven into saturation current by a positive pulse from the trigger circuit. A maximum peak anode current of 3·5A is obtainable from this valve under these conditions. This current can be adjusted by switching two 1Ω resistors in series with the valve filament. A 1Ω potentiometer also in series with a filament gives a fine control of the time sweep providing continuously variable speeds from 3μsec down to about 15mμsec. A diagram of the complete time-base circuit is shown in Fig. 1. The trigger circuit for V_9 is simply a 20A3 thyratron which can be triggered by a positive pulse applied to its first grid via a 470pF capacitor. The grid can also be negatively biased up to 500V from a 250kΩ wirewound potentiometer, connected across the negative supply. This potentiometer is mounted on the front panel of the oscilloscope since it is useful in adjusting the delay of the time-base trigger. This variable delay is very convenient particularly when viewing a small portion of a pulse, and reduces the length of the delay cable required. The maximum obtainable delay by means of this time bias method depends on the initial rise time of the trigger pulse. Since this rise time must be less than about 0·1μsec in order to keep time jitter at less than 1mμsec, the variable delay cannot be much greater than about 0·1μsec.

When thyratron V_8 conducts, it discharges a twelve section artificial line into its characteristic impedance of 1kΩ connected to the cathode of the thyratron. The line is charged to +500V through a 470kΩ resistor producing a 250V square pulse across the 1kΩ cathode resistor. This pulse is applied via socket A to the grid of the cathode-ray tube (to brighten-in the trace) and to the grid of the time sweep valve V_9.

The Beam Modulating and Cathode-ray Tube Protection Circuit

A diagram of the beam modulating circuit is shown in Fig. 2. The circuit was constructed to the manufacturers recommendations for the protection of the cathode-ray tube. It has been suggested that the grid should not go negative by more than 200V with respect to the cathode. This has been prevented by connecting a QS95/10 stabilizer valve (V_3) between the cathode and grid with a 1kΩ resistor. Owing to the rectifying action of this stabilizer valve it was not possible to obtain complete suppression of the beam near the end of the flyback period. A D1 diode V_2 was connected across the input trigger between capacitors C_4 and C_8 to the cathode-ray tube grid in order to prevent negative pulses reaching V_3.

The Power Supply Circuit

The 8kV negative e.h.t. supply to the cathode-ray tube was produced by a 150kc/s power oscillator of the reverse feedback, tuned anode type. The output from the secondary winding (L_8) of the oscillator transformer was rectified, by diode V_6 (see Fig. 3). A second diode V_7 is used to provide a positive 8kV supply which serves as the anode voltage of V_9 the time sweep valve when reduced to 3kV by a suitable resistive potential divider. The oscillator valve V_5, was a 6P28 beam tetrode power output valve fed from a 500V positive d.c. supply with the primary of the r.f. transformer connected in the anode. The primary is tuned by the parallel capacitor (0·02μF) C_{30}. Sufficient feedback was delivered to the grid of V_5 to maintain oscillations by means of a small metal sphere 3/16in diameter placed near the secondary of the oscillator transformer T_3 with a grid leak of 100kΩ. The complete circuit diagram of the generator is shown in Fig. 3.

General Arrangement and Screening of the Circuits

In an attempt to eliminate any interference from external sources with the time-base and the beam modulating circuits of the cathode-ray oscilloscope, the design of screening and layout of the various circuits was carefully considered. The cathode-ray tube together with the time-base circuit and controls were mounted on one chassis and the power unit on another so that they could be mounted one above the other in a standard P.O. rack. A thin layer of copper was evaporated on to the outer surface of the cathode-ray tube. This copper film was connected to earth in order to help minimize interference. Around this copper screen was placed a 1/16in thick Mumetal screen which was terminated one inch from the insulated base of the cathode-ray tube. In addition both the input leads to the Y plates and the complete time-base circuit were built inside separate 1/16in thick copper boxes. The time-base valves were mounted on a small sub-chassis inside the box. Connexions to the Y plates were brought out through the side walls of the box with coaxial connectors. To complete the screening a cylindrical shield made of copper was placed around the

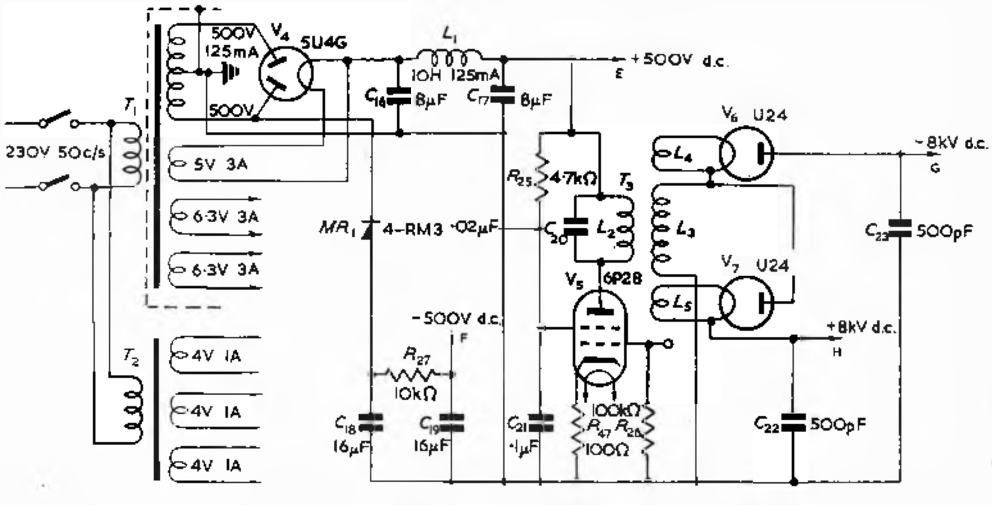


Fig. 3. E.H.T. supply circuit

T_3 R.F. TRANSFORMER

Primary (L_2) *One section 300 turns of 10/0·0028 litz wire wound on 1in dia. former.*
Secondary 1 (L_3) *Six sections of 800 turns of 10/0·0028 litz wire on 1in dia. former.*
Secondary 2 (L_4) *Two turns of 22 gauge d.c.c. wound round live end of Secondary 1 and insulated for 8kV working.*
Secondary 3 (L_5) *One turn of 22 gauge d.c.c. wound round earthy end of Secondary 1 and insulated for 8kV working.*

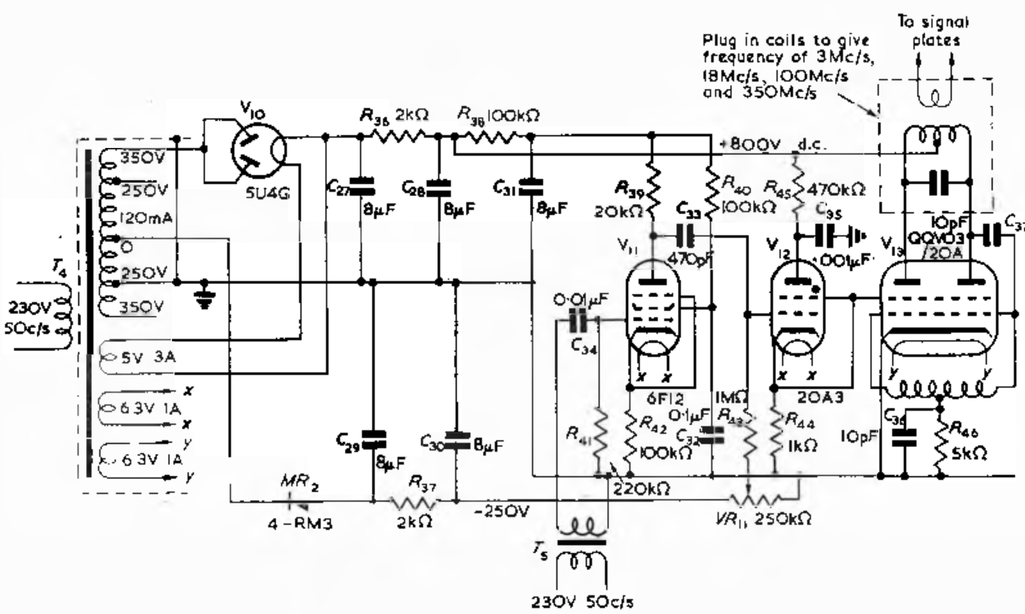


Fig. 4. Calibrating oscillator circuit T_5 = *peaker transformer*

parts of the beam modulating circuit which were mounted on the base of the cathode-ray tube.

The Calibrating Oscillator

In order to calibrate the various time sweeps of the oscillograph it is necessary to have an oscillator with frequencies up to about 300Mc/s and with sufficient voltage to give 1 or 2cm deflexion on the cathode-ray oscilloscope screen. For repetitive work in which the cathode-ray oscilloscope and pulse generator are triggered at 50 times per second, the time jitter of the oscillator must be less than 1mμsec. The usual method for frequencies up to 20Mc/s is to shock excite an *LC* circuit with a suitable pulse. At frequencies of a few hundred megacycles per second this method is impracticable. Several methods were tried and the most reliable circuit obtained is shown in Fig. 4.

The oscillator valve V_{13} is a double tetrode QQV03/20A, and has its control grid connected to a four turn coil of 12 s.w.g. copper wire, the centre point of which is earthed through a 10pF capacitor and a 5kΩ carbon resistor connected in parallel. The anodes are connected to suitable r.f. oscillator coils. The screen grids, which are internally connected together, are driven positive by a 200V pulse developed across the cathode resistor of a 20A3 thyratron. The valve is made to conduct rapidly and oscillates with a frequency controlled by the inductance and capacitance between the two anodes.

For repetitive operation at 50c/s the 20A3 thyratron may be triggered from a peaking transformer or other suitable source. When, however, the initiating pulse had a comparatively slowly rising front (as in the repetitive case) there was considerable jitter of the calibration wave on the screen. This was eliminated by introducing a pulse sharpening stage V_{11} and an *RC* differentiating network between the peaker output and the thyratron trigger. Fig. 5(c) shows a typical 325Mc/s calibration wave on a 15mμsec time sweep. The thickening of the trace is due to the combined jitter of the oscillator and time-base. This is estimated to be less than 0·25mμsec. The frequency of the oscillator was checked and adjusted by means of a wavemeter loosely coupled to the oscillator coils. With the type of time-base used here it is essential to calibrate after almost every observation since variations in the filament current of the 11E3 due to mains fluctuations cause changes in the time sweep.

The Camera

The camera built for use with this cathode-ray oscilloscope incorporates a Wray $f/1$ bloomed lens with a focal length of 2in and was designed to use 35mm perforated film. The camera was mounted on a viewing cone which clamped on to the front panel surrounding the cathode-ray tube face. The film cassettes which hold 25ft of film consist of two cylinders which can be rotated one inside the other. In the open position the slots in both cylinders coincide. This design avoids the inevitable piece of black velvet and the consequent scratching of the film. The film is wound on by a sprocket wheel which is operated from the outside of the camera case.

Performance

The performance of the instrument is perhaps best illustrated by some oscillograms. Fig. 5(a) shows an 0·05μsec long single shot pulse obtained by discharging a length of coaxial cable into its characteristic impedance by means of a hydrogen thyratron. In Fig. 5(b) the front of a similar pulse is shown on a 15mμsec time-base for a pulse repetition

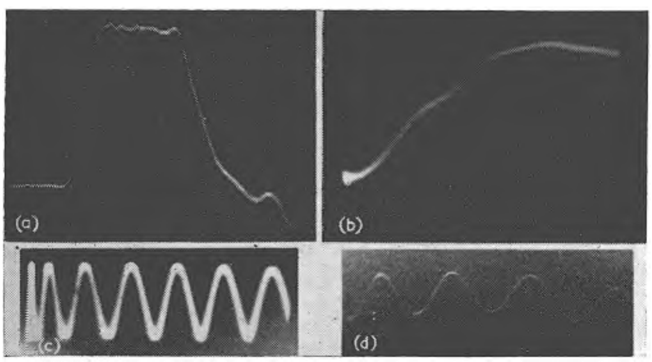


Fig. 5. Oscillograms

(a) 0·05μsec pulse single shot.
(b) Leading edge of an 0·05μsec pulse. 50 cycle repetition, 1sec exposure.
(c) 325Mc/s repetitive calibration wave, 2sec exposure.
(d) 350Mc/s single shot calibration wave, negative intensified.

rate of 50 times per second, exposure time 1sec. These photographs plainly indicate that with this oscilloscope it is possible not only to view single shot pulses of the order of 50mμsec long but also the combined jitter of a hydrogen thyratron and the time-base is less than 1mμsec when used repetitively.

The variation of the sweep speed on a logarithmic scale, is plotted against the filament voltage of the 11E3 in Fig. 6. The graph is almost linear up to 4V and then flattens off and remains constant up to 6V. The shape of this curve is similar to the one given by Smith for a 3D21 valve, except that in this case the curve is shifted towards the lower voltage scale and consequently the saturation region occurs at a lower voltage.

The linearity of the time sweep for different values of the filament voltage of the 11E3 was quite good except for the beginning of the trace on the faster sweeps. This region of non-linearity lasts for less than 10mμsec and is most likely due to the sweep valve not becoming immediately fully conducting. This region can be avoided by delaying the brightening pulse for about 10mμsec or by delaying the signal pulse so that it occurs on the more linear part of the trace.

In certain classes of work it is necessary to examine single stroke transients and, since the electron beam only traverses the c.r.o. screen once, the amount of light produced is so small that special photographic techniques have to be used. One factor in the recording of such fast transients, namely the design of the camera, has already been dealt with, but an investigation into the various types of commercially available films and developers, which are suitable for this type of work was made. In general the conventional high speed orthochromatic films such as Kodak, R60 and Ilford 5G91, were found to be as good as high speed Ilford HPS, panchromatic films. In fact in many instances the orthochromatic film was found to be superior, particularly when intensification of the negative was necessary. Several types of developers were tried but good results were obtained using a general purpose MQ developer. With the orthochromatic films a certain amount of over-development could be tolerated, but with panchromatic films over-development caused fogging. The negatives

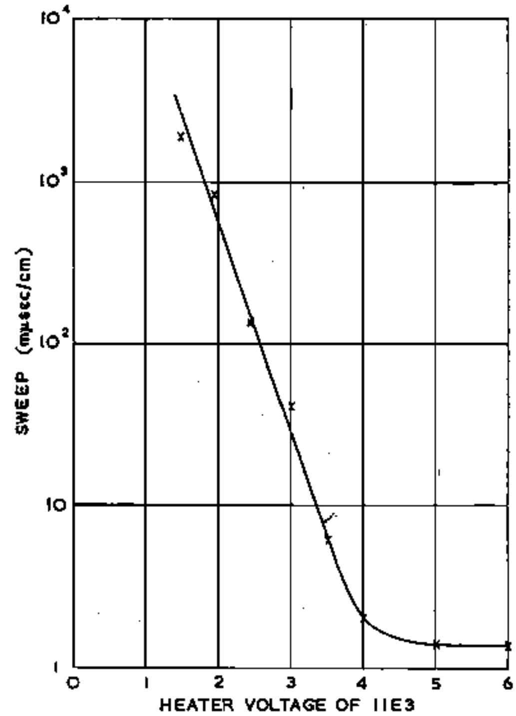


Fig. 6. Relationship of sweep to filament voltage for an 11 E 3 valve

produced when recording very high speed transients are hardly dense enough to make good prints although they are good enough to enable measurements to be made. In order to make good prints the negatives have to be intensified. A good intensifier suitable for this purpose was the quinone-thiosulphate type, suggested by Muehler and Crabtree[6]. Apart from the initial hardening of the negative, this is a single solution intensifier. A print using an intensified negative of a single shot trace of 350Mc/s oscillation is shown in Fig. 5(d).

Acknowledgments

The authors wish to thank Mr. B. G. Churcher, Manager of Research Department, and Dr. Willis-Jackson, F.R.S., Director of Research for permission to publish this article.

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Waveguide Power Dividers

By T. G. Hame*

This article deals with the development of a new type of wideband power divider designed for use in aerial feeding systems. Four types are discussed, 1:1 and 1:2 ratio T-junctions, 1:1 H-plane Y-junction and a combination of the 1:2 ratio T-junction and the Y-junction to provide a 1:1:1 ratio power divider.

In many aerial systems, suitable radiation patterns are produced by feeding several waveguide elements with predetermined proportions of the total input power. It is, therefore, necessary to provide some form of power divider which will produce the necessary power ratios over the operating frequency band of the aerial elements.

The simplest type of power divider consists of a rectangular waveguide operating in an H_{01} mode with a thin conducting plate inserted parallel to the H-field as shown in Fig. 1. If the plate is infinitesimally thin, the electromagnetic field in the waveguide is not affected and there is no change in the input impedance Z. As the field is uniform in the y direction, the voltages which appear between the conducting plate and the waveguide walls are in

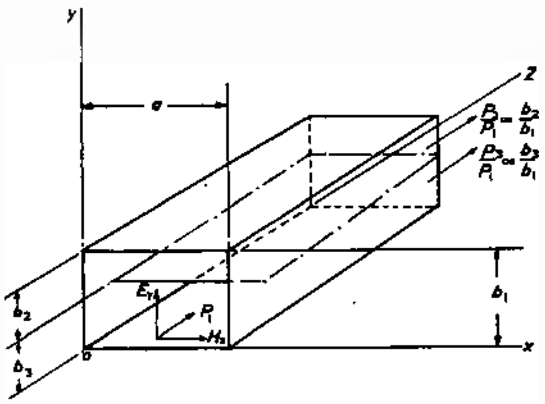


Fig. 1. Simple power divider

the ratio of $Vb_3/Vb_2 = b_3/b_2$. The H-field is also unchanged, therefore the currents in the two sections will be the same and hence the power flowing in each section will be directly proportional to b_2 and b_3. The characteristic impedance of the sections has, of course, been reduced by a factor of b_2/b_1 or b_3/b_1.

If one now branches the power divider into a series T-junction, as shown in Fig. 2, and makes the colinear arms 2 and 3 half the thickness, or impedance, of the series arm waveguide then, for a power division ratio of 1:1 or $b_2 = b_3$, the input impedance Z_1 to the series arm is equal to

$$Z_{o2} + Z_{o3} = (Z_{o1}/2) + (Z_{o1}/2) = Z_{o1}$$

if the discontinuities resulting from the right-angle bends are ignored. The discontinuities can be removed by inserting a mitre at each corner of the junction as shown in Fig. 3, so that Z_1 does, in fact, equal Z_{o1}. If the dividing plate is now removed, the field inside the waveguide will remain unchanged and the two mitres (back to back) provide the necessary power division and impedance matching.

Two experimental power dividers of this form, having power division ratios of 1 : 1 and 1 : 2 respectively, are described. The last section of this article is devoted to some experiments with an H-plane symmetrical Y-junction which, in conjunction with a 1:2 ratio power divider, forms a 3-way divider having a power division ratio of 1:1:1. The use of a resonant iris to obtain an improvement in the bandwidth characteristic of the Y-junction is also described. Experiments were made in X-band using WG15 and half-WG15 guides.

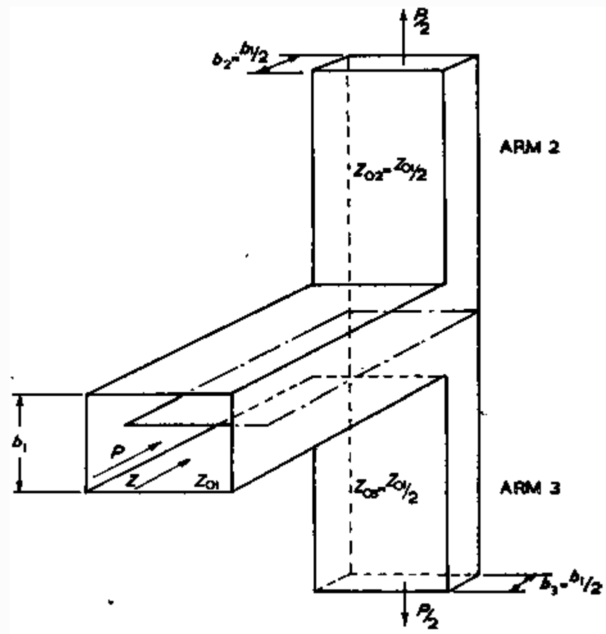


Fig. 2. Series T-junction

Double-Mitred T-Junction Power Dividers

Power Ratio 1:1

A drawing of the experimental divider is shown in Fig. 4. It was constructed entirely in brass, with the double mitre mounted on the centre line of the series arm waveguide to obtain a 1:1 ratio. The double mitre could be removed quite easily by unscrewing the two retaining screws. The following measurements were made to obtain the height (h) of the double mitre for the optimum input standing wave ratio to the series arm over a 500Mc/s band. All other dimensions were kept constant during this operation, and the two colinear arms were terminated in matched loads. The height (h) of the double mitre was initially the full half-guide width and was reduced in small steps. Fig. 5 shows the standing wave characteristics obtained in the vicinity of the optimum value $0{\cdot}1488\lambda$ of the height h.

Power Ratio 1:2

To obtain the required 1:2 ratio of power, an experimental divider was constructed as for the 1:1 divider, but the ratio of b_2 to b_3 was altered to 1:2 (Fig. 6). The main guide remains matched as far as the power division is concerned, provided the impedances of the colinear arms are similarly altered, for any position of the dividing plane. But, for reasons to be shown later, these outgoing branches

* *E.M.I. Electronics Ltd.*

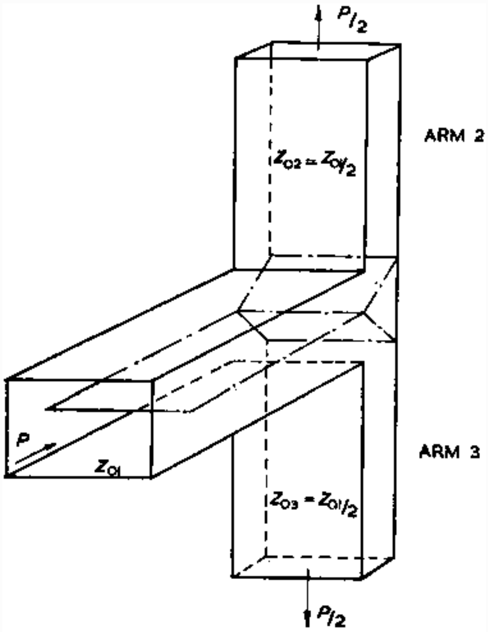


Fig. 3. Insertion of mitres

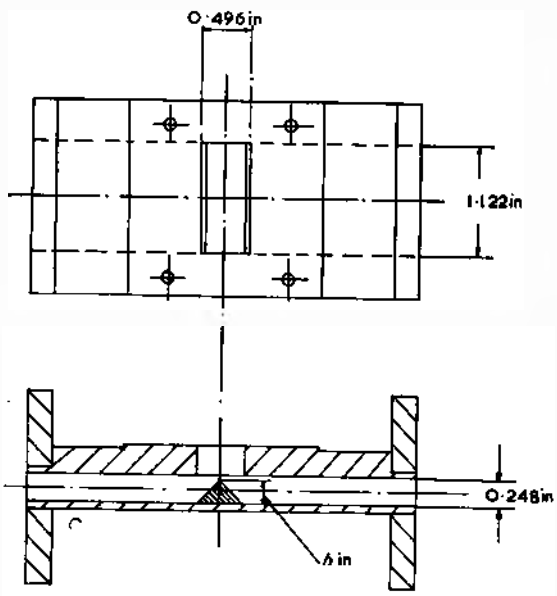
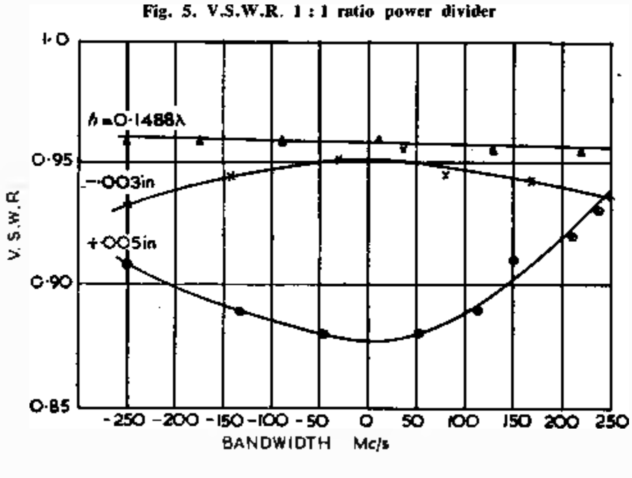


Fig. 4. An experimental divider

Fig. 5. V.S.W.R. 1 : 1 ratio power divider



were made equal to half the impedance of the input guide. The mitre therefore has the double problem of determining the division ratio and, at the same time, matching the two input fractional guides into the half guides.

A symmetrical double mitre was retained and the height (h) again varied in small steps. The power division ratios were also measured for each dimension h. Fig. 7 shows the

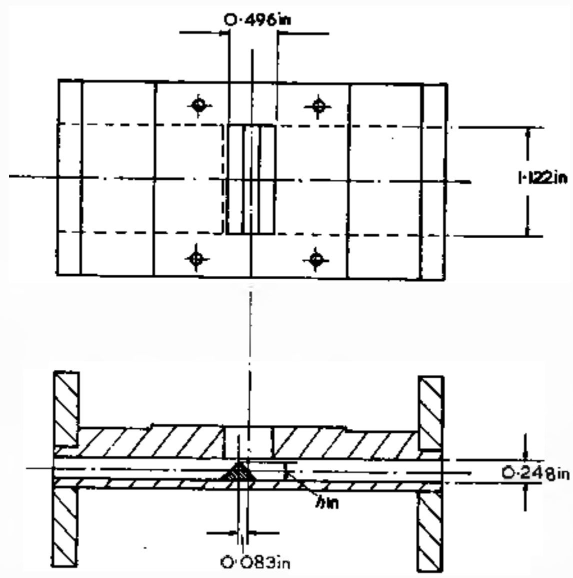


Fig. 6. Experimental divider for 1 : 2 power ratio

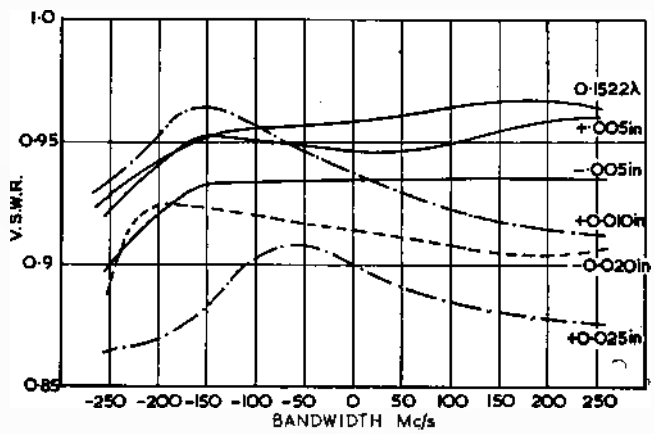


Fig. 7. V.S.W.R. 1 : 2 ratio power divider

input standing wave ratio to the series arm over a 500Mc/s band for various values of h, while Fig. 8 indicates the variation in power ratio for different values of h.

H-Plane Y-Junction Power Divider

Ratio 1 : 1

Another useful type of power divider is the Y-junction and, for reasons which will be apparent from a perusal of the final section of this article, an H-plane symmetrical junction was selected for investigation.

If power is incident in arm A of the Y-junction shown in Fig. 9 some of this power will be equally divided between arms B and C, while the remaining portion will be reflected back to the source in arm A due to the impedance mismatch and also to the junction effect. Considering the impedance mismatch, arms B and C will appear in shunt with one another and present a terminating impedance of $Z_0/2$ to arm A which has an impedance of Z_0. As Z_0 is real, the standing wave ratio looking into arm A will be

$$\frac{Z_o/2}{Z_o} = 0{\cdot}5$$

An experimental check of this result gives an average figure of 0·54 for the input standing wave which indicates that the junction effect is small compared with the impedance mismatch for a symmetrical junction.

An impedance measurement at arm A also indicates that, if a capacitive post 0·101λ in height is inserted in the junction at 1·866λ from flange A, a near-unity match is obtained over the operating band. In practice, however, the optimum position of the post was found to be 1·860λ from flange A and, with the post in the new position, an input standing wave ratio of 0·99 was obtained at 50Mc/s off the centre frequency as shown in Fig. 10. To check whether this position is, in fact, the optimum, the post was moved by a distance $\lambda g/2$ towards the generator and a second curve of v.s.w.r. versus frequency was plotted. This showed a maximum value at the centre of the band, which indicates

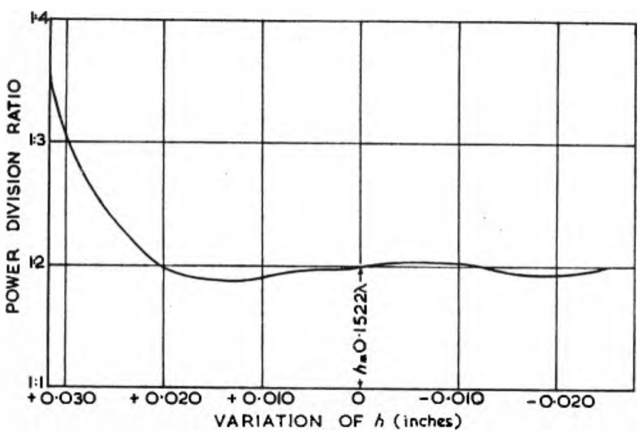
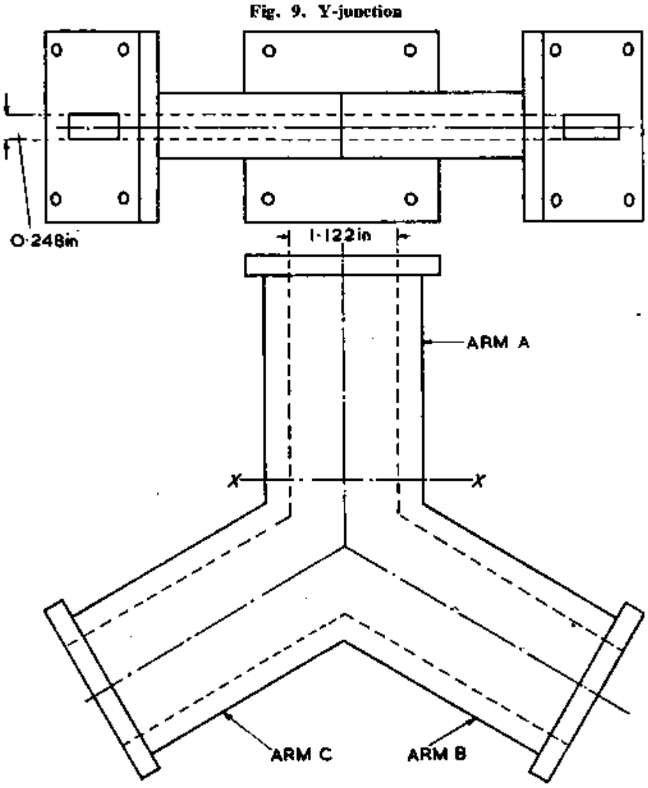


Fig. 8. Power division ratio 1 : 2 power divider

Fig. 9. Y-junction



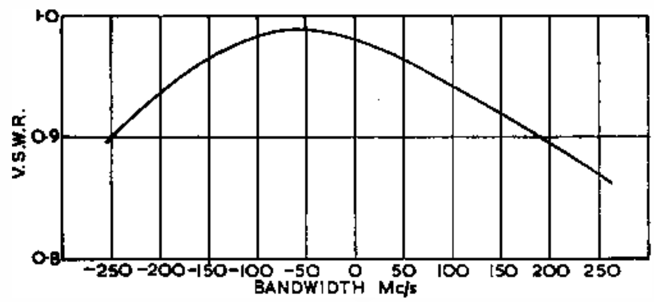


Fig. 10. V.S.W.R. Y junction with post in arm A

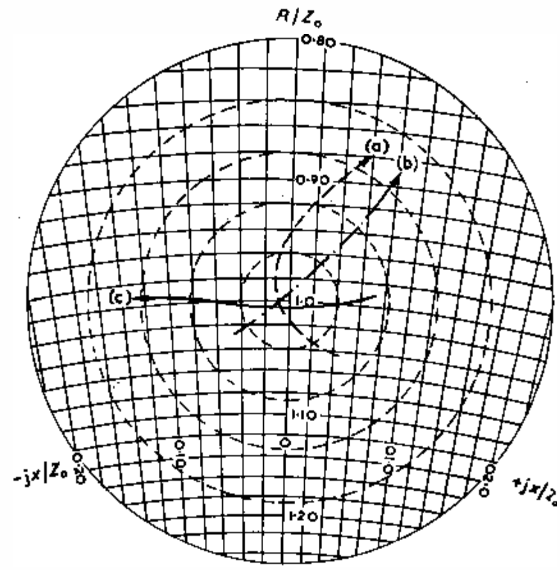


Fig. 11. Admittance of post-matched Y-junction
Key: (a) Flange A; (b) Junction xx (Fig. 9); (c) 1·412 from Flange A

that the optimum dimension has been obtained. The slight shift between the frequencies at which the optimum values were obtained is most probably due to the first post being located in the junction at a place where the electromagnetic field is complex, while the second is in a region which is fairly undisturbed. The power division ratio and the input standing wave ratio to arm B were also measured and, within experimental error, the power division was 1 : 1 and the input standing wave ratio to arm B was approximately constant at 0·4 over the required band.

Use of a Resonant Iris

The best method of obtaining a good frequency/v.s.w.r. characteristic with a waveguide device which has discontinuities, is, of course, to insert another device which has the opposite frequency characteristic to that of the discontinuity. One convenient method of achieving this is to insert a resonant iris.

Admittance curves taken at different planes in arm A of the post-matched Y-junction are shown in Fig. 11. It was necessary to find a plane in arm A at which the admittance curve lies along the line of unit conductance, as this is the plane at which the resonant iris must be situated. This plane was found to be 1·412λ from the flange. The approximate dimensions of the iris were obtained by Slater and Huxley's[1] method, but the final dimensions were determined experimentally. The iris, which was 0·030in thick was placed in a length of waveguide attached to standing wave measuring gear and terminated in a matched load. The admittance in the plane of the iris was measured and the dimensions of the inner rectangle modified until the iris was resonant at approximately the centre frequency (Fig. 12). The iris was then removed from the waveguide and inserted at the predetermined position of the Y-junction (Fig. 13) and the input admittance was measured. The results are shown in Figs. 14 and 15 which indicate considerable improvement in the input standing wave ratio.

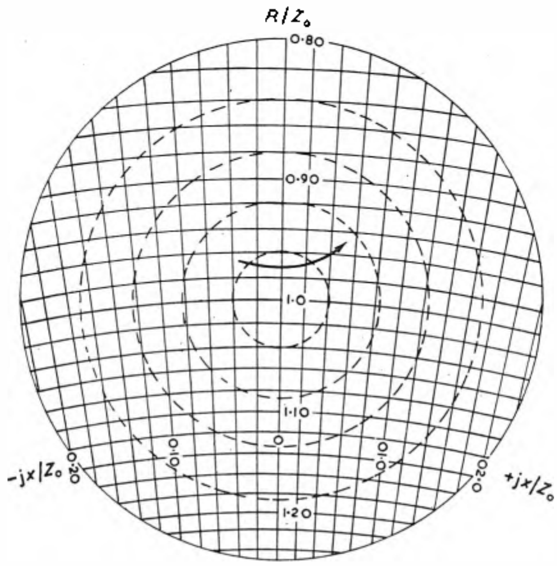


Fig. 12. Admittance of 0·030in copper

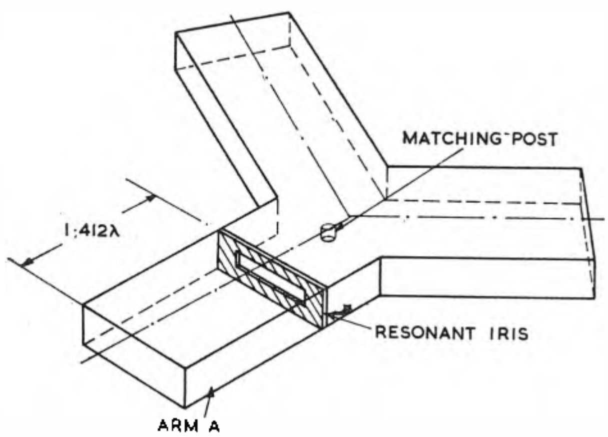


Fig. 13. Insertion of iris in divider

Two further thicknesses of resonant iris were also investigated in a similar manner, the thicknesses being 17 s.w.g. (0·056in) and 12 s.w.g. (0·104in) respectively, both being made of brass. The results obtained are shown in Figs. 15(b) and 15(c) respectively.

Three-way Power Divider (Ratio 1:1:1)

If the Y-junction is combined with the double mitred 1 : 2 ratio power divider, a compact 3-way divider with a 1 : 1 : 1 ratio can be formed as shown in Fig. 16. If one unit of power enters the series arm of the T-junction then $\frac{1}{3}$ is sent directly out of arm C while the remaining $\frac{2}{3}$ is split equally between arms A and B of the Y-junction. Thus, if three matched aerials are being fed, each will receive an equal amount of power.

Two superimposed, but not interconnected, power dividers were constructed to feed two sets of three aerials, one set being used for transmission and the other for reception (Fig. 17). The dimensions used were those obtained from the previous experiments but no resonant iris was included in the Y-junction, for the sake of simplicity.

The input admittances of the transmitter and receiver arms are shown in Fig. 18, the six aerials being simulated by matched loads for the purpose of measurement. The

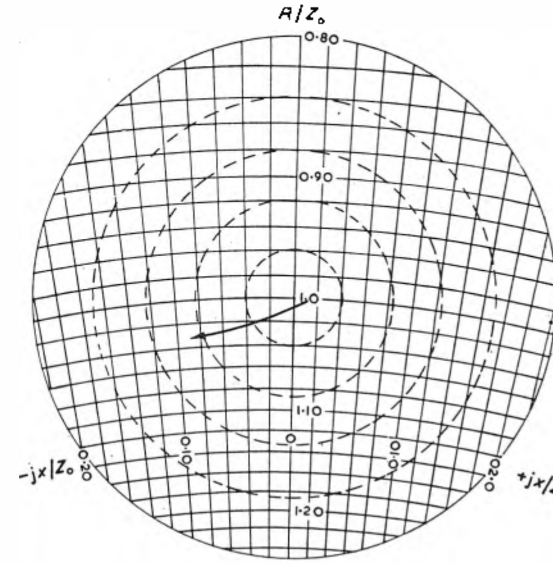


Fig. 14. Admittance of Y-junction with post and 0·030in iris 1·412in from flange A

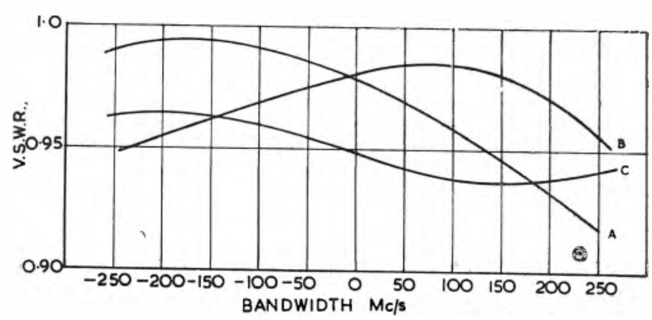


Fig. 15. (a) v.s.w.r. Y-junction with post and 0·030 copper iris
(b) v.s.w.r. Y-junction with post and 17 s.w.g. iris
(c) v.s.w.r. Y-junction with post and 12 s.w.g. iris

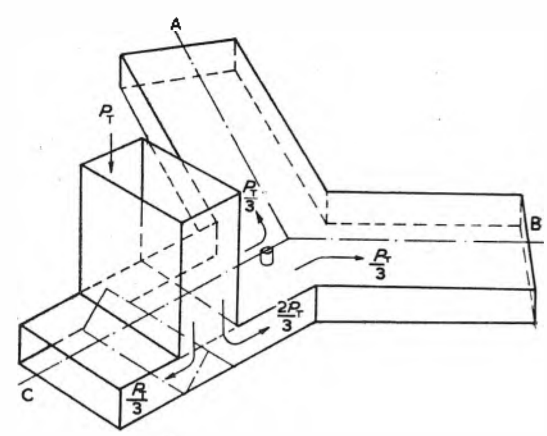


Fig. 16. Compact three-way divider

transmitter power division over the frequency band was measured with a thermistor bridge arranged to replace each matched load in turn and the received power was measured by attaching the thermistor bridge to the receiving arm and supplying power successively to each aerial input, the other two aerial inputs being loaded. The results of these measurements are shown in Figs. 19 and 20 respectively.

Due to the close proximity of the double mitre and the Y-junction some interaction was expected. It was decided, therefore, to adjust the height of the matching post in the transmitting Y-junction to see what improvement could be effected in the input admittance. Fig. 21 shows the input admittance for slightly different heights of post and indicates the existence of an optimum value at a height of 0·1085λ where a v.s.w.r. of better than 0·86 can be obtained over a 1 000Mc/s band.

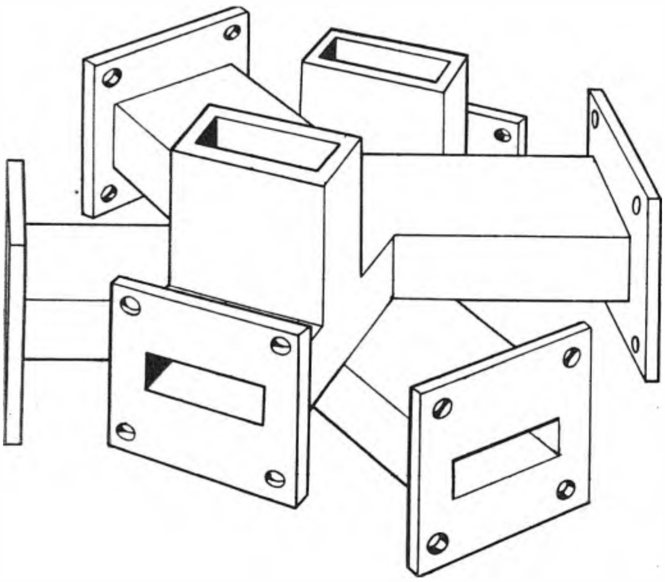


Fig. 17. Divider for feeding two sets of three aerials

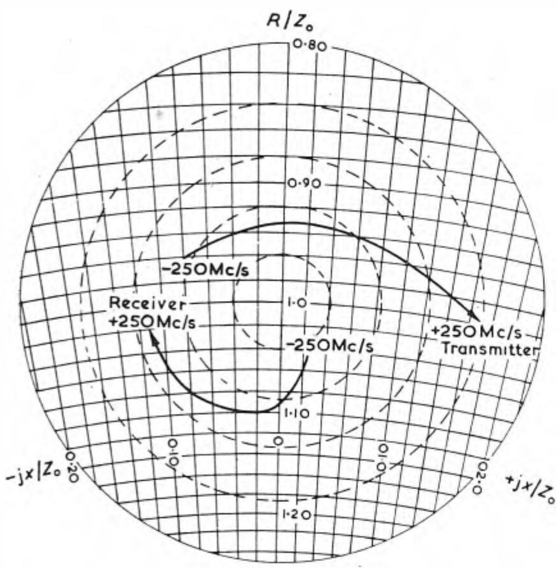


Fig. 18. Input admittance of transmitter and receiver arm of 3-way power divider

Conclusions

With the optimum height h of the double mitre, for the 1:1 divider the input v.s.w.r. is very nearly constant at 0·96 over the required 500Mc/s band, while a tolerance of ±0·002in causes only a small drop in the mean v.s.w.r. to drop to approximately 0·94 at the extreme edges, or at the centre of the band.

For the 1:2 ratio divider, the optimum height (h) gives a mean v.s.w.r. of 0·955 over the band. A tolerance of ±0·002in causes only a small drop in the mean v.s.w.r. to 0·95. It can also be seen that, provided the mean input v.s.w.r. is better than 0·925, a 1:2 power division is obtained.

It is possible that the standing wave ratio could be improved by re-shaping the mitre, but this was not investigated.

Both junctions are, therefore, suitable for wideband power dividers, the differences between the optimum dimension of h being exceedingly small for a change in power

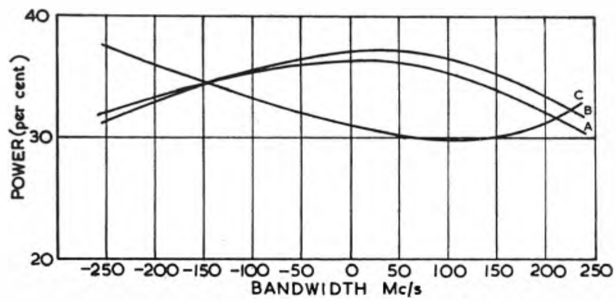


Fig. 19. Percentage power in each arm of 3-way power divider

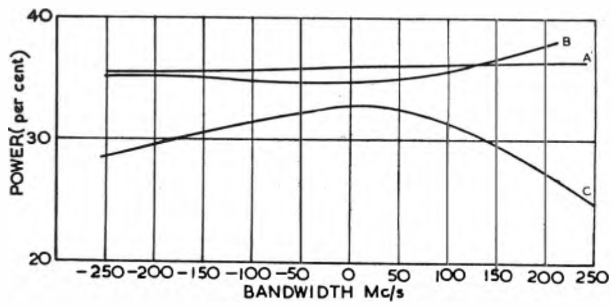


Fig. 20. Percentage power developed at receiver of 3-way power divider

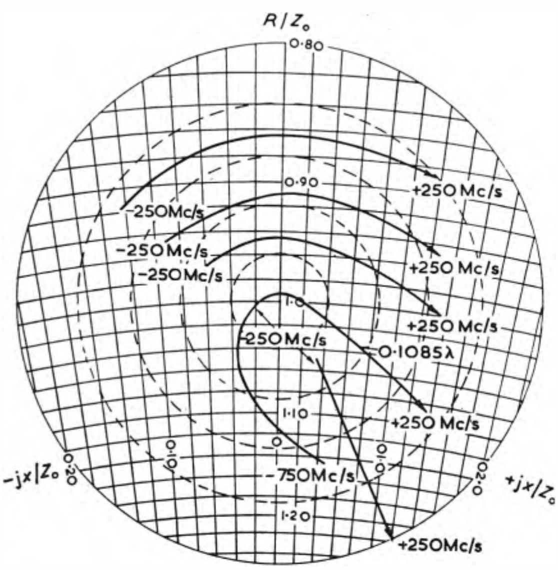


Fig. 21. Input admittance of transmitter arm with various heights of post

division ratio of 1:1 to 1:2. The dividers also have the following further advantages.

Ease of manufacture and assembly.

Extreme compactness (no matching elements required in the series arm).

Reasonable mechanical tolerances.

The post-matched Y-junction, when used as a 1 : 1 power divider, will give a fairly good performance over a 500Mc/s band; the input standing wave ratio drops to 0·87 at the edge of the band. In many simple waveguide systems, this performance would be quite adequate. If, however, a resonant iris is added, an input standing wave ratio of better than 0·95 can be obtained with a properly located 17 s.w.g. iris.

The double-mitred, Y-junction power divider forms a compact assembly for feeding three aerials, or other devices. A 1 : 1 : 1 ratio of power division and a very satisfactory input standing wave ratio are maintained over a wide band with reasonable mechanical tolerances. The differences between the input admittance of the transmitter and receiver arms (Fig. 18) is due to the distance between the Y-junction and the double-mitred T-junction being different in the two cases.

Acknowledgments

The author wishes to express his thanks to Messrs. P. R. Cloud, G. Harris, J. Thraves and A. C. V. Baker for their invaluable assistance during the development of the dividers and to E.M.I. Electronics Ltd. for permission to publish this article.

REFERENCE

[1] HUXLEY, L. G. H.: The Principles and Practice of Waveguides (Cambridge University Press 1947).

A Direct Recording Cathode-Ray Tube*

Various attempts have been made to record direct from a cathode-ray tube without having recourse to a photographic process. With the type of cathode-ray tube screen to be described it is possible to make direct recordings on for example, a 'Teleteldos' type paper.

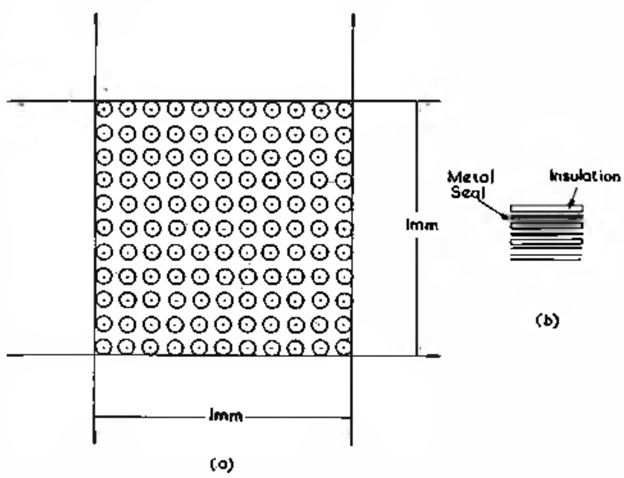


Fig. 1 (above). Front view and cross-section of screen

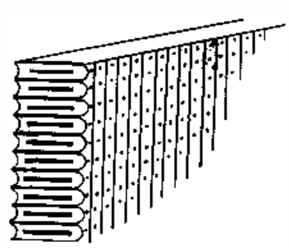
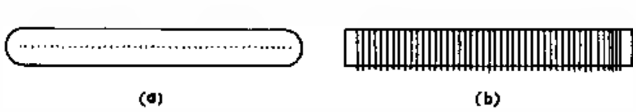


Fig. 2 (right). Section of screen

Fig. 3 (below). Linear arrangement for recording tube



The screen consists, in essence, of a flat glass plate through which a number of wires are passed to form a mosaic, (see Figs. 1 and 2). The number, or density of wires can be varied up to a maximum of about 100/mm^2. The screen can be ground and polished to any desired thickness and the glass to wire seals are adequate to support the degree of vacuum normally used in a cathode-ray tube. The construction of such a screen is the subject of recent British Patents†.

Cathode-ray tubes on this principle but with the wires in a straight line (not a mosaic) have already been constructed with 100 points in 10cm (see Fig. 3), and successful tests have been carried out on them.

It is possible that a screen of this type may have other uses than in a cathode-ray tube designed for recording, as for example, storage tubes and pick-up tubes.

* *A communication from H. F. Charlotte.*

† *Charlotte, H. F. British Provisional Patents 36995/55 and 36996/55.*

A Mine Telephone

Mechanization at pit bottom and other underground areas results in greatly increased noise, and it is consequently necessary to introduce some form of speech reinforcement for the conventional underground telephone system at important points such as tipplers, loading stations, transfer stations, and near the coal face. The BTH "Portavox" amplifier and loudspeaker have been developed specially for this purpose and are now in production.

The equipment is installed in addition to the mine telephone, and serves to amplify and reproduce through a loudspeaker the signals normally reproduced by the telephone earpiece, so that communication can be established under very difficult conditions.

The equipment is offered as intrinsically safe and flameproof; certificate number IS.1066 and FLP.3832/A having been granted by the Ministry.

In operation, the caller rings the bell of the telephone in the usual way and speaks into the mouthpiece. The speech is reproduced by the loudspeaker at the reception point where the "Portavox" is installed, but when the earpiece is lifted from the hook of the receiving telephone so that a reply can be given through the mouthpiece, the amplifier is cut off (thus avoiding 'ring-round') and the signals are reproduced in the normal manner by the earpiece. Loudspeaker communication can thus be achieved over the normal telephone circuit by holding the earpiece hook down when receiving a message and releasing it when speaking into the mouthpiece.

The loudspeaker assembly is part of the intrinsically safe equipment and consequently special fire-resisting material—resin-bonded glass fibre laminate with suitably adjusted ingredients—has been used for the re-entrant type horn, which will neither support combustion nor produce thermite reaction on impact.

Perhaps the most important feature of the "Portavox" system is its ready application to party-line circuits, where code ringing is a necessary but nevertheless tiresome and sometimes lengthy method of making contact with a desired person. "Portavox" enables the message to be announced if the code bell remains unanswered.

Furthermore, circumstances may arise which by their emergency make it imperative to give immediate verbal warning to operators. Instructions of this type can be readily broadcast at high volume by the "Portavox" system.

It will be appreciated that the "Portavox" equipment can also be installed, if so desired, at any point in the line between the normal telephone positions, thus giving greater coverage than the existing system.

A VOLTAGE STABILIZER PRINCIPLE

By C. Billington*, B.A., and E. Chakanovskis*, Dipl. Ing.

The development of a number of voltage stabilizer circuits employing positive feedback is discussed, together with their merits and limitations. The performance, ease of construction, and wide range of output voltages are sufficient for a variety of purposes where moderately high stability (0·02 per cent) and cheapness are paramount considerations.

WITH the greatly increased use of systems employing feedback, it is now readily appreciated that the performance of a degenerative voltage stabilizer is closely related to the loop gain. At first sight the maximum gain attainable is limited by the high tension available for the amplifier and by the number of amplifying stages used. Taking the amplifier high tension from the unstabilized side of the series valve introduces difficulties which have been studied by numerous authors; they have been only partially solved by resort to compensation methods which have somewhat unfortunately been called 'feed forward'. Because there is no closed loop associated with compensation, complicated arrangements are required to keep the compensation critically adjusted over a wide working range. Many commercial stabilized supplies depend upon the device of introducing an auxiliary negative stabilized supply of fixed voltage connected to the negative terminal of the main supply. The most negative line of the system is now used for the amplifier of the positive supply. Only by this technique is it possible to control the positive supply right down to zero output voltage. There are, however, many applications needing a well stabilized high tension supply where the provision of an additional negative supply is uneconomical, and where the facility for adjusting the output voltage to near zero is not wanted. The alternative of increasing the loop gain by increasing the number of stages involves intricate circuits to prevent instability. Particularly when the stabilizers are to be built infrequently by research personnel rather than daily by wirers on an assembly line, it is undesirable for the circuit to include a variety of capacitors and resistors to provide subtle degenerative loops at high frequencies.

Positive Feedback

One of the ways in which the loop gain can be increased without recourse to the methods indicated above is by the use of positive feedback within the loop. Valley and Wallman[1] mention the use of positive feedback in connexion with voltage stabilizers, but the circuit given as an example does not in fact use positive feedback. Increasing the gain by positive feedback makes little demand on high tension voltage, and need present no coupling problems. Some method of adjustment is needed, but it proves less sensitive than the compensation arrangements. The impedance of the stabilizer will be expected to rise more rapidly at high frequencies than without positive feedback. The first arrangement tried is shown in Fig. 1. When in optimum adjustment this circuit restricted output voltage fluctuations to 0·01 per cent over a current range of 20 to 40mA and an input voltage change from 360 to 400V, which corresponds to about ±5 per cent mains change. At a fixed current, say 30mA, an input voltage range of 350 to 430 could be accommodated; this corresponds to ±10 per cent mains change. When no attempt was made to obtain correct feedback adjustment, the performance figure was reduced to about 0·03 per cent over the same ranges. Apart from the performance limitations already implicit, there are a number of other deficiencies:

(a) The amplifier has a dynamic range of only 10V within which 0·01 per cent stability is maintained.

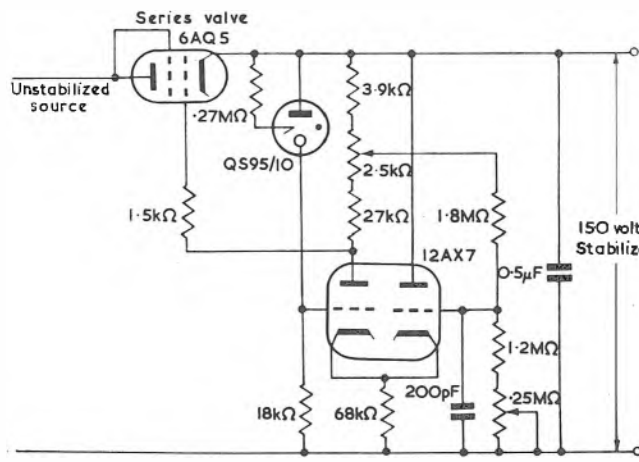


Fig. 1. Elementary stabilizer using positive feedback

Furthermore this 10V range corresponds to a series valve grid bias swing from −10 to −20V. The impossibility of using grid bias values smaller than −10V with this circuit increases the dissipation of power in the series valve and limits its range of operation.

(b) The circuit does not have very good drift characteristics due largely to the high value resistors in the voltage divider for the positive feedback grid of the 12AX7. Carbon resistors had to be used since wire-wound components of the required values were not easily obtainable.

(c) The voltage reference valve is not very aptly chosen. Connecting the ignition electrode to the stabilized side makes for erratic striking, while connexion to the unstabilized side introduces compensation of uncertain magnitude. Voltage reference valves with only two electrodes are more suitable.

(d) The adjustments of output voltage and feedback are not independent.

Emphasis upon amplifier design has led the authors to a rather different conception of stabilizer performance. The performance has two separate features: the degree of stability, and the input voltage tolerance and load current range over which this stability is maintained. The action

* *Commonwealth Scientific and Industrial Research Organization, Melbourne, Australia*

of positive feedback is to keep the amplifier gain high over a certain range of output voltage but outside this range the augmented gain falls off rapidly. Positive feedback thus defines the limits of dynamic range of the amplifier more sharply. At the same time the gain is such as to give a stability approaching 0·01 per cent within the range. In this study a stability of 0·01 per cent has been the target since it appears to correspond with the limit of drift stability using wire-wound resistors in suitable places but not attempting any thermal compensation. The range over which this stability is maintained is almost entirely dependent upon the properties of the series valve. The suitability of various series valve types is discussed in another article[2].

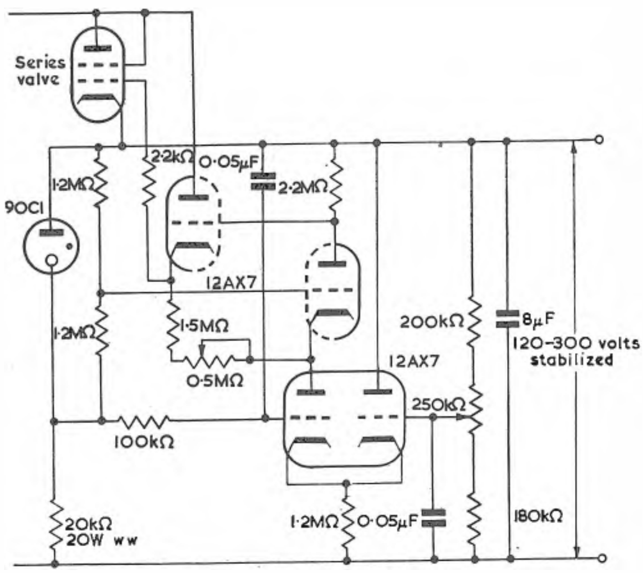


Fig. 2. Stabilizer using two amplifying valves and positive feedback

Cascode Amplifier with Positive Feedback

The next stage of the development of the amplifier is shown in Fig. 2. Two double triodes are used, but with the same effective high tension voltage as before. A bridge containing the voltage reference source in one arm is unbalanced when the stabilizer output voltage changes, and so provides the error signal in conventional feedback analysis. This error signal is amplified differentially by the first double triode. Capacitors improve the sensitivity of the bridge for a.c. signals. Half of the second double triode is in cascode connexion with the appropriate side of the differential amplifier, while the other half is used for positive feedback. In this cascode application there is no difficulty in providing a steady source of potential for the upper grid. Attree[3] describes a stabilizer using a modified cascode arrangement but without positive feedback. By connecting the cathode-follower in at the middle of the cascode instead of down to the negative line, a positive feedback loop is established. Critical adjustment is made by altering the cathode-follower resistance until changes of current in the upper half of the cascode are compensated by current change in the cathode-follower. In fact the net current change in the common cathode resistor of the differential stage is of the order 2µA for a change of 25V at the cascode anode. The arrangement is naturally unstable unless the negative loop is also in operation as shown in the figure. The feedback can be critically adjusted by altering the input voltage or load current or both, and noting the effect on the stabilized output with some form of backed off voltmeter.

The advantages of this two valve system over the one valve system of Fig. 1, are:

(a) A larger dynamic range at the grid of the series valve is available.

(b) By taking the amplifier output from the cathode-follower which is at a higher potential than the cascode anode, the range of the stabilizer extends right down to zero bias on the grid of the series valve.

(c) The stabilized output voltage is negligibly affected by adjustment of the feedback control.

(d) No additional degenerative loops are necessary to prevent high frequency instability.

This circuit has been used at 150V output in conjunction with a 325V transformer, 6X4 rectifier, and a type N37 as series valve. Twenty-four such units were made without any selection of components for a small computer, and all meet the stability specification of 0·02 per cent for ±10 per cent mains variations and load current between 25 and 60mA. Some selection of 12AX7's is necessary so that the effect of heater unbalance in the differential stage does not exceed the tolerance upon output voltage. A small proportion of 12AX7's do not give adequate service in the cascode and feedback position. Four of these units have operated for over 500 hours during the development of other equipment without valve failure or deterioration of performance.

Bench tests have been made with the stabilizer set to a variety of output voltages between 120 and 300V. Although exhaustive tests were not possible, the performance was of the same order at all values. A less satisfactory feature of this stabilizer for variable output voltage is the wide range of current that has to pass through the reference tube, and also the cathode-follower half of the differential amplifier. The type 90C1 was the only valve available which would accommodate so large a current ratio without running too close to the manufacturer's limits, but it is not designed with a view to high stability as a reference source.

Modification for Low Drift and Linear Voltage Adjustment

For another application where low drift was important further study of this principle was made. In the arrangement of Fig. 2 sources of drift were found to lie particularly in the reference source and the carbon potentiometer used for voltage adjustment. The reference valve was changed to a type QS83/3 and operated near the design condition, and the potentiometer was removed in an arrangement giving fixed output voltage. Then by subjecting each resistor in turn to a jet of cold air and noting the corresponding change of output voltage, it was found that the carbon resistors used in the bridge were likely to be largely responsible for thermal drift. Accordingly wire-wound resistors were used in the bridge circuit and were mounted close together. Under these conditions the drift could be reduced to about 0·01 per cent per hour for most of the 12AX7's used in the differential amplifier stage and provided there were no sustained changes of heater voltage greater than about 5 per cent. When the stabilizer was working remotely from the unstabilized source the drift was not affected by load changes within the working range. In stabilized supply units where the stabilizer was situated close to the transformer and rectifier the output voltage would drift about 0·03 per cent after a large change in load current.

The feature of variable output voltage which was abandoned in the interests of low drift can be restored by exploiting the constant current properties of the amplifier. It has already been mentioned that the total variation of amplifier current over the working range is small on account of the action of the positive feedback loop. The percentage change in current through the bridge and amplifier combined is about three times the percentage change in bridge voltage over the working range: the exact ratio is determined by the amount of positive feedback. If this combined current, I, is passed through a 'foot' resistance, R, to the common negative line, the stabilized output voltage is increased above the bridge voltage by an amount IR. By using a linear wire-wound rheostat for the foot resistance, the drift is not impaired, and the output voltage control will have a uniform calibration. Provided the total output voltage is not more than about twice the bridge voltage the performance will be reduced by a factor less than two. However for a given value of R the amount of positive feedback can be adjusted to give optimum compensation for the whole output voltage. If R is to be adjustable some loss of performance is incurred at the extremes of the voltage range, and this will set a practical limit to the magnitude of the range available by this means.

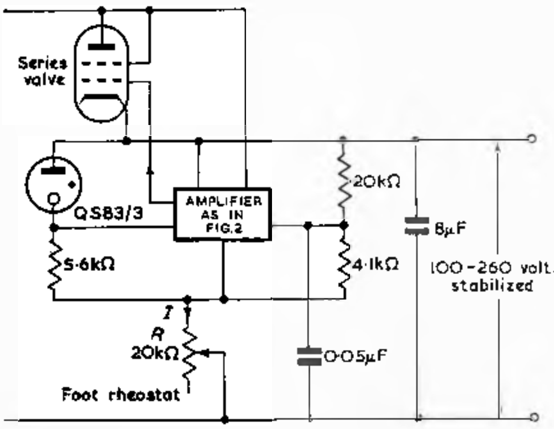


Fig. 3. Stabilizer with low drift and linear voltage adjustment

It was recently found that this method of adjustment has already been published by Amey, Krantz, Clark, and Williams[4] in a general account of stabilized power supplies for instrument applications. So little pertinent detail is given, however, that the claims made are difficult to assess.

Fig. 3 gives details of a stabilizer with resistance values which allow an output range of 100 to 260V. The high tension voltage used by the amplifier is seen to be little more than the burning voltage of the reference valve, and is independent of the output voltage. The convenient limit of output voltage range was found to be about 200V, though it is not to be implied that this can be achieved with a fixed unstabilized source voltage. As Amey and his co-workers point out, voltage reference valves can be placed in series with the foot rheostat to obtain higher output voltages. This has been tried with a single type QS1200 to give a voltage range of 250 to 400V with excellent results; but to use twelve of these valves, or twenty type 85A2, in order to reach 2 000V seems to be an uneconomical method of stabilizing in this voltage region.

With the positive feedback adjusted for optimum performance at mid range (180V) in the circuit of Fig. 3, the stability at this voltage is normally better than 0·02 per cent. If the output voltage is then set to 100V, or 260V, only 0·05 per cent stability can be achieved. The maximum output resistance is less than 0·1Ω up to 2kc/s, and then rises to 1Ω at 20kc/s. The residual ripple is less than 300μV. These values apply to about 80 per cent of the 12AX7's used. The majority of valves which did not give satisfactory performance were known to have had unfortunate mechanical histories. 'Reliable' types have not been available to the authors, and with this percentage of good valves they were not concerned to investigate failures. There must be some selection to obtain satisfactory heater balance, if the performance is not to be spoiled through the effect of heater voltage changes. An unselected valve is likely to change the output voltage by 0·05 per cent for a 10 per cent mains change. Heater compensation has been tried using half a 12AX7 as a diode with adjustable load in a conventional manner. While sufficient compensation can easily be achieved, it is much harder to avoid a transient following a step change of heater voltage, on account of the different relaxation times of the heaters even though the same valve is used. The transient only amounts to about 0·02 per cent which may be tolerable for some purposes.

85V Version

In an attempt to explore briefly the minimum voltage attainable by this principle, a stabilizer was built to give a minimum output voltage of 85V, i.e. 2V higher than the reference source voltage. The stability at 85V could be made 0·02 per cent over the working range, but the current regulation of the bridge was not good enough to allow the voltage to be increased more than about 20V by a foot rheostat without serious loss of performance.

Comment on Instability

Of the many units that were constructed during the course of this development, none has shown any tendency to oscillate provided that proper attention is paid to grid stopper resistors for the series valve. Considerable latitude in layout is permissible. With miniature valves screen grid stoppers have been found unnecessary, but when a number are used in parallel, each valve should have a separate control grid stopper of at least 1·5kΩ. Most of the designs have used types N37 and EL80 with up to four valves in parallel. The type 12E1 requires more careful and individual attention to stoppers especially when several valves are used in parallel.

The method of positive feedback has not found ready acceptance so far in the literature partly because it is generally harder to prevent oscillation, and partly because the optimum amount of positive feedback is often too critical. The foregoing amplifier circuit in which the feedback setting is not unduly critical appears to us to have three necessary features:

(a) High intrinsic gain (without positive feedback).

(b) A minimum of components in the positive loop.

(c) Very small dynamic range of input with respect to the positive line.

Conclusion

The stabilizer described can deliver output voltages as low as 85V with stability about 0·02 per cent without the use of an auxiliary negative stabilized supply. The high gain necessary for such a stability is achieved by the use of positive feedback, but the frequency response is not unduly limited (0·1Ω at 2kc/s), and the layout is not critical. The output voltage is varied by a linearly calibrated control.

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A Discussion of Series Valves for Small D.C. Voltage Stabilizers

By C. Billington*, B.A.

A graphical method of assessing the performance of a series valve is presented. On this basis fifteen valve types are compared, and some valves not normally used in this application are shown to have unsuspected advantages.

MANY authors have contributed to the literature upon degenerative voltage stabilizers since the monumental paper of Hunt and Hickman[1] in 1939. Almost all are concerned with the arrangement of the amplifier, and methods for compensating its shortcomings. The present author is unaware of any published consideration of the role of the series valve and the proper choice of valve. Trigg[2] has offered an analytical treatment of a simple stabilizer, but he is concerned with the whole circuit and does not specifically mention the series valve type used. Types 2A3, 807, 12E1, 6AS7, 6V6 (6AQ5), 6AU5, and 6Y6 have been widely used as series valves but reasons for their choice have rarely been stated.

Grid Bias Limits

The essentials of the circuit to be considered are shown in Fig. 1. The voltage reference valve and the amplifier are operated from the stabilized side of the series valve. Referred to the stabilized positive line as datum, the output voltage of the amplifier corresponds to the grid bias on the series valve. Consequently the dynamic range of the amplifier fixes the limits of bias voltage that can be impressed on the grid of the series valve. In order that the degree of stabilization should be maintained over the whole working range of input voltage and load current, the gain of the amplifier must remain sufficiently high over its dynamic range. Thus although it might appear at first sight that distortion in the amplifier is of little account because of negative feedback, in fact distortion will set a limit to the dynamic range if a uniform stabilizer performance is to be maintained. It is not intended to discuss amplifier performance in this article, except to establish that there is a limit to the grid bias available for controlling the series valve. The method, which will be described, of assessing the capabilities of any series valve requires only that the dynamic range of the associated amplifier be known. In stabilizer circuits[3] developed concurrently with the present investigation, the use of positive feedback in the amplifiers made their output voltage limits more obvious. For the sake of numerical definiteness in tabulation, the grid bias limits 0 and −25V have been chosen.

For amplifiers connected as in Fig. 1 entirely on the stabilized side of the series valve, zero bias will be beyond the undistorted limit since it corresponds to cut-off conditions in the final amplifying stage. It is desirable to have the limit at zero bias if possible, since this is the condition of minimum power dissipation for a given series valve. The arrangement of Fig. 1 can be improved in practice by using a cathode-follower between the amplifier output and the series valve grid. The anode of the cathode-follower has to be connected to the unstabilized supply. The cathode-follower output is a few volts more positive than the amplifier anode output, and consequently the zero grid bias condition of the series valve may lie within the useful working range of the amplifier. The additional connexion to the unstabilized side of the series valve reduces the stabilization ratio from

$$1 : 1/G\mu_s$$

to

$$1 : (1/\mu_s + 1/\mu_{cf})/G,$$

where G is the amplifier gain, and μ_s and μ_{cf} the amplification factors of the series valve and cathode-follower respectively. Thus provided a high-μ triode is used for the cathode-follower the stabilization ratio is not much affected.

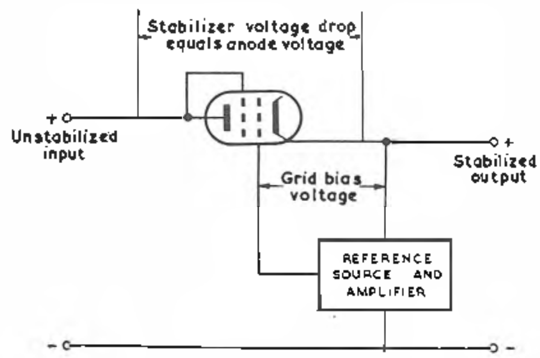

Fig. 1. Basic arrangement of stabilizer

Graphical Method

The working range of any series valve can be represented by a permitted area on a plot of its anode characteristics. This area is bounded by (a) zero grid bias, since the amplifier cannot supply grid current; (b) rated maximum anode dissipation; (c) rated maximum anode or screen voltage; (d) maximum grid bias that can be supplied by the amplifier (taken as −25V for the purposes of later numerical analysis). The area is shown hatched in Fig. 2, but for simplicity the vertical boundary corresponding to (c) has been assumed to lie outside the area. For all points within this permissible area the stabilizer responds in the same way to changes of load or input voltage: both correspond to changes in grid bias of the series valve.

In practice the stabilized source voltage tolerance must be substantial if ±10 per cent mains variations are to be accommodated. For stabilizers with fixed output voltages up to 300V, ±50V will cover ±10 per cent mains variations, but to allow more latitude in an unstabilized source design employing stock transformers ±70V may be more suitable. Peak-to-peak ripple voltage remaining after filtering must be added to the maximum tolerance of input voltage caused by mains variations to arrive at the gross input voltage tolerance the stabilizer has to accommodate. If a variable current drain is required from the stabilizer the working point will also move up and down in the diagram of Fig. 2. Because of the finite internal conductance of the

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unstabilized source its voltage will rise as the current is reduced. Neglecting mains variations for the present, the working point will move along a line of slope equal to the internal conductance of the source. With a capacitive load to rectifier this line is not quite straight but curves towards higher voltages at very low currents; this curvature has been been neglected for simplicity in the remainder of the discussion. Now taking account of mains variation at the same time as current variation, the working point will be seen to lie within a parallelogram whose horizontal side is equal to the gross input voltage tolerance. The slope of the other pair of sides is given by the internal conductance of the unstabilized source. This parallelogram is the limit of the usuable working area and is shown cross-hatched in the figure. It must, of course, lie wholly within the permissible working area, and this restriction determines the length of the sloping side. The top horizontal side must fit between the zero-bias boundary and the maximum anode dissipation boundary, and the lower right-hand corner lies on the maximum bias boundary. The projection of the parallelogram on the current axis determines the maximum and minimum currents that may be drawn from the series valve for a given gross input voltage tolerance. This is not to say that stabilization will necessarily fail beyond these current limits: failure will only occur if at the same time the input voltage is too near the appropriate extreme value.

Maximum Rating of Valves

Those boundaries of the permitted working area of Fig. 2 which are curves of constant bias are immediate limitations to proper stabilization. The boundaries corresponding to maximum ratings, however, do not affect the degree of stabilization until the series valve fails either from overheating or electrostatic breakdown. In many laboratory instruments valve failure is so rare, and then so often due to heater failure, that it might be economical to exceed the maximum ratings in some cases. For instance, the maximum anode dissipation curve might well be made to pass through the point in Fig. 2 corresponding to the source design centre voltage plus 5 per cent instead of plus 10 per cent. Maximum anode and screen voltage ratings fall into rather a different category. With 6AQ5 and N37, two of the types which have been used most during the investigation, the permissible working area would be greatly reduced if the maximum screen voltage ratings were adhered to strictly. Since no failures of these valves have occurred during nearly two years' investigation and use in equipment, it is thought that in the case of these two types at least there is little disadvantage in exceeding the maximum voltage ratings by as much as 100V. There is certainly much to

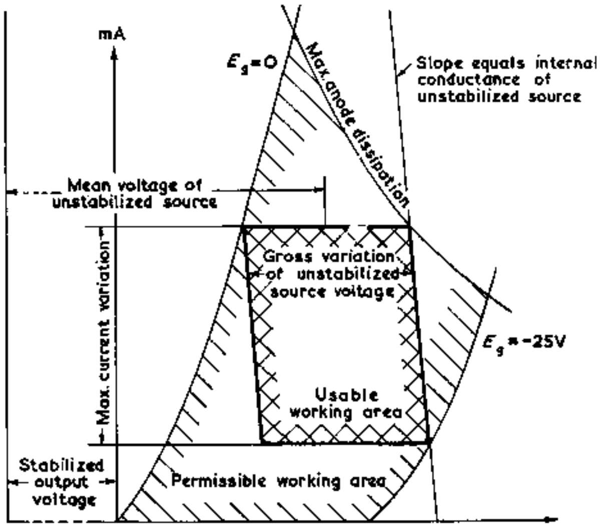

Fig. 2. The relation between the anode characteristics of a series valve and the range of operation of a stabilizer using that valve

be gained from their increased performance range. An article by Knight[4] on valve reliability is valuable in this connexion although he is primarily concerned with maximum possible reliability. In the performance data given in Table 1, the manufacturer's maximum dissipation ratings have been regarded as absolute ratings, but the maximum voltage ratings of some types have been somewhat exceeded.

TABLE 1

Series Valve Performance Data

	GROSS INPUT VOLTAGE TOLERANCE AT MAXIMUM CURRENT DRAIN				±50V			±70V		
	VALVE TYPE	AMPLIFICATION FACTOR	ANODE DISSIPATION (W)	ANODE RESISTANCE AT ZERO BIAS (Ω)	MEAN ANODE VOLTAGE AT MAX. CURRENT (V)	MAXIMUM CURRENT (mA)	MINIMUM CURRENT (mA)	MEAN ANODE VOLTAGE AT MAX. CURRENT (V)	MAXIMUM CURRENT (mA)	MINIMUM CURRENT (mA)
Miniature	EL84	18	14	1300	183	62	1	185	55	1
	N78	24	12	1800	180	52	1	188	46	1
	EL83	24	11	2000	175	49	1	183	44	1
	EL80	22	10·4	1800	160	48	1	170	42	1
	6CL6	23	9	1500	143	46	1	156	41	1
	N37	10	12	1000	127	66	11	139	57	19
	6AQ5	9½	14	2100	193	57	18	201	52	24
	EL81*	7	10	770	108	64	29	121	53	36
Octal and Others	EL38*	16	33	1000	195	136	2	205	122	2
	EL34*	11	33	1000	183	144	8	190	128	16
	807	7½	25	1400	205	98	22	215	88	37
	6AU5*	6	10	770	110	66	34	120	54	46
	12E1*	5	40	440	150	198	152	—	—	—
	6Y6	5	14	670	125	80	60	—	—	—
	2A3	4·2	15	670	135	85	75	—	—	—
	6AS7	2	26	280	—	—	—	—	—	—

NOTES.—1. Grid bias limits, 0 to −25 V.
2. All tetrodes used in triode connexion.
3. Figures for types derived from tetrode characteristics.*

Input Filtering

If a conventional d.c. source using a transformer and rectifier is to be stabilized there is an economical limit to the reduction of residual ripple by input filtering. One volt r.m.s. corresponds to less than 1 per cent mains change at the anode of the series valve of a supply giving 150V stabilized output. The residual a.c. appearing at the stabilized output which is normally less than 1mV r.m.s. is almost entirely due to hum from the amplifier and pick-up from the power transformer, and is not increased by considerably increasing the value of the ripple at the anode of the series valve. However, much more than one volt of ripple at input will seriously increase the gross input voltage tolerance that has to be accommodated by the stabilizer, and so in practice will reduce the range of output current that can be supplied.

Application of Actual Valve Types

Diagrams of the kind illustrated in Fig. 2 have been constructed for fifteen valve types, and two values of gross input voltage tolerance, 100 and 140V. The mean voltage drop across the series valve at maximum current, and the extreme values of operating current have been tabulated for either case. Other relevant valve data have been included. The internal resistance of the unstabilized source has been regarded as fixed at 1kΩ. An asterisk denotes that triode characteristic curves have not been available to the author, and have been estimated from tetrode characteristics. Accordingly figures for these valves will be less reliable. The list includes valves which have been widely used, and also some which would appear to merit wider use in this application. Types N37, EL80, and 6AQ5 have received most attention in practical circuits.

Assessment

It would be difficult to assign a general figure of merit that covers all purposes for which this kind of stabilizer is used. The miniature valves naturally have smaller maximum currents, but may be used in parallel without much circuit complication, and will often save space and increase the flexibility of layout. In Fig. 3 a comparison of eight of the valves is made against total power dissipation. The extremes of current are taken from Table 1 for a gross input voltage tolerance of 100V in order that more of the more usual valves could be compared. Total power dissipation is derived by multiplying the extreme current by the mean anode voltage, and adding the heater power. The extreme points are joined by straight lines whose length and proximity to the left-hand side of the figure give some indication of merit.

The valves may be classified approximately into three groups according to amplification factor which appears to be of first order importance. There is little overlapping between the groups, but this may only be due to the small sample of valve types tabulated.

Amplification Factor = 16 and Above

These types are suitable for general purpose stabilized supplies which are likely to be used at negligible current drain. Most of these types will stand considerably more than 140V gross input tolerance, and so will allow in addition a measure of variation of the stabilized output voltage without altering the input voltage. Provided that the amplification factor is at least 16 the minimum current is practically zero, and the maximum current is closely related to the maximum rated anode dissipation; no great advantage accrues from higher amplification factor values since the rated maximum anode voltage becomes the limiting factor. Most of this group show up poorly in the matter of power dissipation (Fig. 3).

Amplification Factor = 7 to 16

Types in this intermediate group are useful in applications where the minimum current drain is not negligible, but the current range is still substantial. The current range increases with amplification factor. There is a considerable spread of mean voltage drop across the series valve. Comparing types N37 and 6AQ5 which are the same physical size, it is evident that the smaller anode resistance of the former more than offsets its lower anode dissipation, and gives measurably better performance.

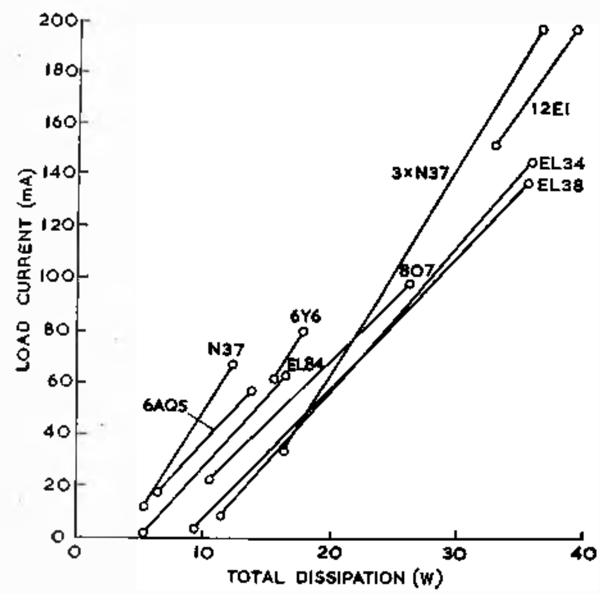

Fig. 3. Comparison of a number of series valves in stabilizers accommodating 100V input voltage tolerance

Amplification Factor = 6 or Less

This group cannot be recommended for use in the restricted conditions of this study. Low anode resistance does not compensate for low amplification factor and the valves require amplifiers of much larger output voltage range.

Conclusion

A graphical method of ready applicability has been tested in practical design, and found to be useful in the selection of valve types for series valve operation in d.c. stabilizers. Amplification factor is shown to be probably the most important single parameter. In particular the type N37 appears from the analysis to be one of the most suitable valves for this operation, although its 13V heater rating can be a disadvantage. The valve operates satisfactorily from two series-connected 6·3V heater windings one point of which should be connected to cathode. It has given excellent service in this establishment. For applications requiring a wider range of load current, or where much variation of the stabilized output voltage is contemplated, the types EL34, EL38, and EL84 deserve consideration. Only the EL80 has as yet been used here for this purpose.

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LACE

(The Luton Analogue Computing Engine)

(Part 2)

Non-Linear Units Used in a General Purpose Analogue Computer

By J. C. Jones*, B.Sc., and D. Readshaw*, B.A.

This part describes:

(a) An electronic multiplier; (b) electronic general purpose function generator, both developed for use in LACE, which equipment was the subject of Part 1.

The accuracy of the multiplier is $\frac{1}{4}$ per cent of full scale output, while its frequency response is flat to 4kc/s. Phase shift is 3° at 300c/s and 45° at 4kc/s.

The function generator, having seven variable 'break points' and eight independently variable slopes, will approximate to any single valued function with the added facility that the corners which occur at its 'break points' can be 'rounded off' by a parabolic curve.

An Electronic Multiplier

The authors, set with the requirement of an electronic multiplier, with the following five specifications,

Accuracy: better than 0·5 per cent of full scale output

Frequency response: 0 to 4kc/s

Phase lag: less than 10° at 1kc/s

Simple to adjust and align

Relatively easy to engineer

felt that existing multipliers known to them at the time all failed to meet such a specification on one count or other.

The mark space multiplier[1] or the Goldberg[2], although having the possibility of attaining the accuracy and frequency response required would not, in the authors' opinion, be as simple to adjust or engineer as the multiplier to be described.

In the authors' experience the McNee crossed field multiplier[3] calls for a complicated and extensive setting up procedure to maintain its accuracy.

The diode quarter squarer[4,5] requires a considerable number of diodes and associated measured resistors to attain an accuracy of 0·5 per cent and once completed is practicably impossible to adjust against ageing resistors and variable diode contact potentials. Little is known of the hyperbolic[6], Hall effect[7] and Kerr cell[8] multipliers, and their operational use was difficult to assess.

The method of multiplication described here is based on the fact that a triangular waveform can be clipped in such a way that the mean level of the resulting waveform is proportional to the square of the voltage controlling the clipping level.

Having acquired then a parabolic function in this manner the identity $(x + y)^2 - (x - y)^2 = 4xy$ can be applied to achieve the required product xy.

The engineering of the multiplier is greatly simplified by the fact that seven identical two-valve d.c. amplifiers are required which are constructed as sub-units on printed boards and plugged on to the main chassis.

In addition to these amplifiers a three valve amplifier and its associated chopper amplifier are made up as sub-units and plugged in. Remaining equipment being four diodes, a bank of twenty 100kΩ resistors and one LC filter. By means of potentiometers brought out to the front panel of the multiplier, setting up procedure consists simply of zero setting the input and output amplifiers before adjusting two bias voltages and finally scaling the output.

The multiplier was in its final stages of development before the authors read of a similar multiplier[9] developed in the U.S.A. and of Mr. Norsworthy's[10] use of the basic principles in this country.

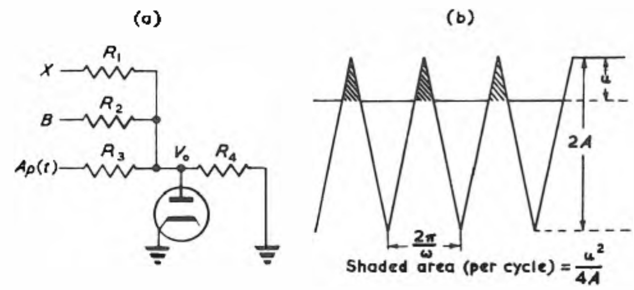


Fig. 10. (a) Basic unit, (b) Triangular waveform clipping (Multiplier)

Principles of Operation

As implied in the introduction the multiplier is based on the identity

$$(x + y)^2 - (x - y)^2 = 4xy$$

The terms of the left-hand side are achieved by clipping a 20kc/s triangular wave at an amplitude determined by the quantities $(x + y)$ and $(x - y)$. The resultant waveform being filtered to give the product xy.

From Fig. 10(b) it can be readily seen that if a triangular wave is clipped such that the amplitude of the resultant waveform (i.e. shaded area) is equal to u then the area per cycle, or d.c. content, of the resultant waveform is proportional to u^2.

The analysis of this is done in Appendix 1, which concludes that for a network such as that shown in Fig. 10(a). For the conditions

$$R_1 = R_2 = R_3 = R_4 \text{ and } B = A \text{ (positive bias)}$$

$$V_{1dc} = -(1/16A)\, X^2 \text{ for } -2A \leqslant X \leqslant 0$$

By reversing the diode, replacing X by Y and letting $B = -A$ (negative bias)

$$V_{1dc} = (1/16A)\, Y^2 \text{ for } 0 \leqslant Y \leqslant 2A$$

Clearly, then, four such networks with inputs $(x + y)$, $-(x + y)$, $(x - y)$ and $-(x - y)$ would yield the equation

$$(V_1 + V_2 + V_3 + V_4)_{dc} = -(1/16A)\,[(x + y)^2 - (x - y)^2]$$
$$= -(1/4A)\, xy$$

* *The English Electric Co. Ltd. Guided Weapons Division.*

for all $A \leqslant x \leqslant A$

$-A \leqslant y \leqslant A$

From this it is seen that one must provide

(a) A triangular waveform whose amplitude must be at least equal to the maximum expected input,

i.e. if $-50 < x < 50\text{V}$

$-50 < y < 50\text{V}$

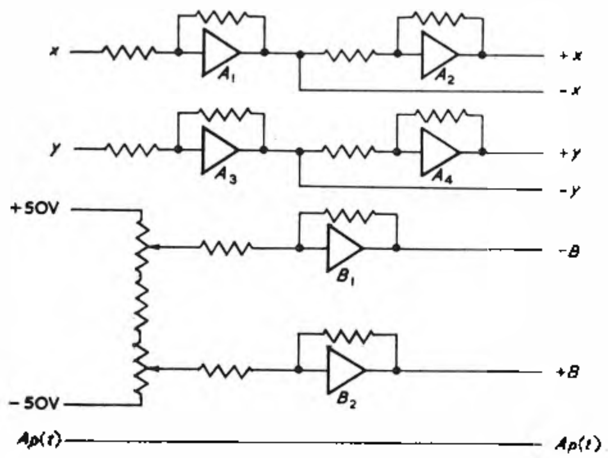


Fig. 11. Input arrangement (Multiplier)

the triangular waveform must be at least 100V peak-to-peak.

(b) Two d.c. voltages equal to the amplitude of the triangular waveform and opposite in sign.

(c) Accurate phase inversion of the inputs x and y.

(d) Accurate summation of the outputs of the four networks V_1, V_2, V_3 and V_4.

(e) An efficient filter to extract the required d.c. voltage.

The triangular wave is generated separately from the multiplier and is fed to two multipliers and two function generators as mentioned in later paragraphs. A closed loop system is used involving a bistable circuit and an integrator. Linearity, although difficult to measure, was such that an accuracy of 0·05 per cent was achieved for squaring a single input using a single network (such as Fig. 1(a)). A frequency of 20kc/s was found to be a convenient balance between efficient clipping and broad computing bandwidth.

The two valve amplifiers (Fig. 11), used to provide from low impedance the bias voltages, the inputs and phase inversion of the inputs and the output product, have an open loop gain of about 500. Tested with their working feedback ratio of 1 : 1 their frequency response was flat up to 40kc/s. Over the range -50 to $+50$V linearity was 0·1 per cent of full scale (i.e. $\pm 0{\cdot}05$V) and they showed very little drift (± 5mV).

The resistance networks used to generate $\pm(x + y)^2$ and $\pm(x - y)^2$ are made up of carefully matched 100$k\Omega$ resistors and summed into a three valve drift compensated d.c. amplifier (Fig. 12). The a.c. amplifier used for drift compensation consists of a double triode and a 50c/s relay. A low pass *LC* filter with a cut-off frequency of 8kc/s achieves efficient attenuation of the 20kc/s carrier and its harmonics. Phase lag at 300c/s is 3°. Attenuation in the pass-band (0 to 4kc/s) is less than 0·1dB.

A three-way switch on the front panel enables inputs from an accurate reference divider to be switched in for the purpose of setting the two bias voltages. It was found sufficient to adjust the biases such that an output was multiplied by four for doubling the inputs, checking, of course, that the output is zero for input zero.

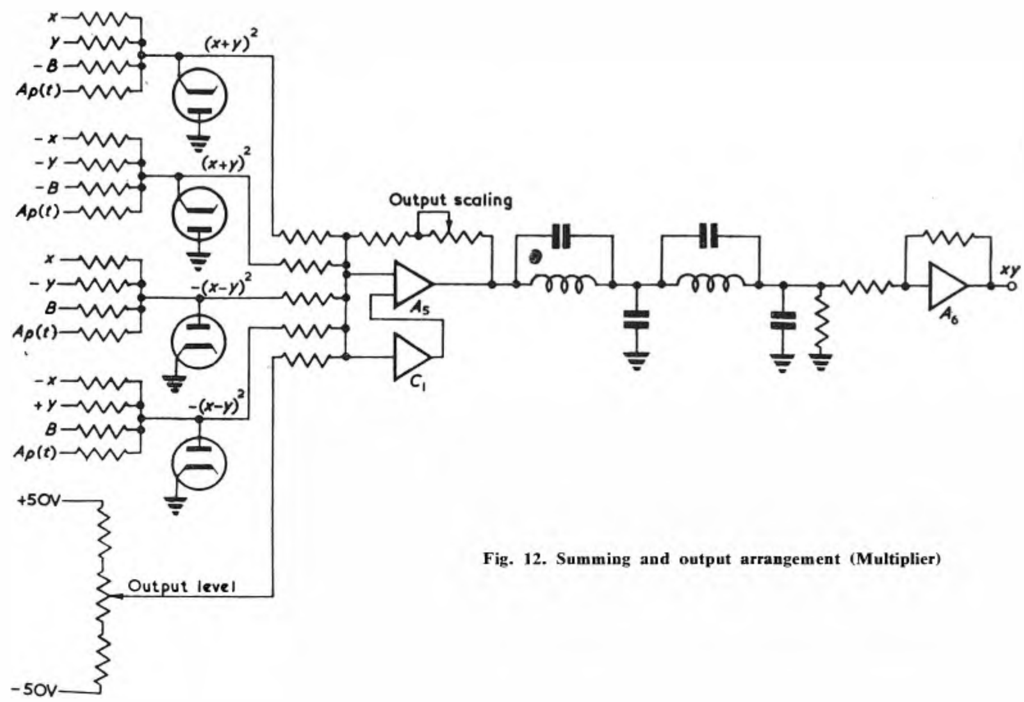


Fig. 12. Summing and output arrangement (Multiplier)

A photograph of the complete unit is shown in Fig. 13.

Practical experience has proved the multiplier to be very simple to adjust from time to time. The low drift figure of the d.c. amplifiers used is such that only the bias adjustments are necessary to maintain the accuracy required.

ACCURACY

Accuracy tests took the form of multiplying two equal inputs for the four variations of sign.

	Input 1	*Input 2*	*Output*
(1)	0 —— +50	0 —— +50	0 —— 50
(2)	0 —— +50	0 —— 50	0 —— −50
(3)	0 —— −50	0 —— −50	0 —— +50
(4)	0 —— −50	0 —— +50	0 —— −50

40 readings were taken, i.e. $|x| = (5, 10 \ldots\ldots 50\text{V})$

The mean of the four full scale outputs was used as a reference full scale and 10 ideal outputs calculated. The 40 errors (ϵ) were then given from

$$\epsilon = \frac{\text{ideal—actual}}{\text{reference full scale}} \times 100 \text{ per cent}$$

It was found then that $|\epsilon|_{max} = 0{\cdot}25$ per cent and $(\Sigma\,|\epsilon|/40) = 0{\cdot}1$ per cent.

Function Generator

It was first required to make a choice of the most suitable form of function generation for incorporation in LACE. Well-known types of function generators include:

(a) Tapped and tapered potentiometers

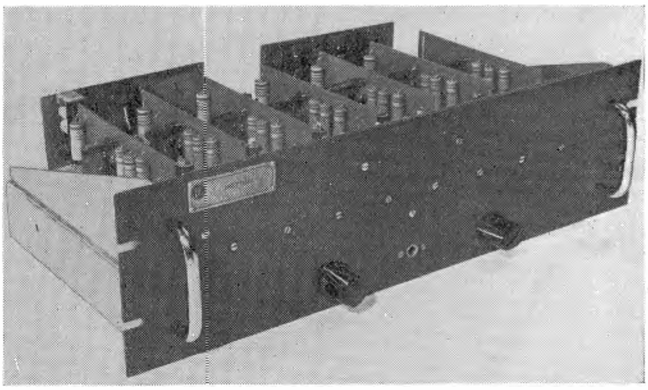


Fig. 13. Front panel view (Multiplier)

(b) Photo-formers

(c) Biased diodes.

Type (a) was considered unsuitable due to the serious limitation on frequency response by inertia of the motor shaft required to drive the potentiometers. Type (b) has an extensive frequency response. However, unless a large library of masks is available, considerable delay is involved in making up a new mask whenever a new function is to be represented. Thus, type (c) was finally chosen, as with this means, a good high frequency response is attainable, together with flexibility and rapidity in setting up a given function. Two models involving different biasing arrangements were investigated and a brief description of model A is given below, followed by a more complete description of model B, which was ultimately chosen as being the more suitable. Both models approximate the given function by a number of linear segments, and it is required to calculate the optimum values of segment slopes and break points between segments, to yield the best approximation.

MODEL A FUNCTION GENERATOR

The basis of this model[4] is a network of three resistors and a diode connected to the input grid of a high gain feedback d.c. amplifier as in Fig. 14. $-B$ is a constant negative voltage source. Then, the relationship between X_o and X_1 is:

$$X_o = 0 \qquad \text{for } X_1 \leqslant B(R_1/R_2)$$

$$X_o = \frac{(X_1R_2 - BR_1)R_o}{R_1R_2 + R_2R_3 + R_3R_1} \text{ for } X_1 > B(R_1/R_2)$$

$$\text{and } \delta X_o/\delta X_1 = \frac{R_2R_o}{R_1R_2 + R_2R_3 + R_3R_1}$$

This gives the characteristic shown in Fig. 15, which consists of two linear segments, one of zero slope and the other dependent on the value of all four resistors. The break point depends on $-B$ and the ratio $R_1 : R_2$.

If a number of these units are now connected to the input of the amplifier, each will contribute a characteristic of the same form, with slopes and break points depending on the resistance combination in each unit. The individual contributions and their resultant will be as shown in Fig. 16. The break points of the complete characteristic are defined

Fig. 14. Basic Diode Unit (Model A Function Generator)

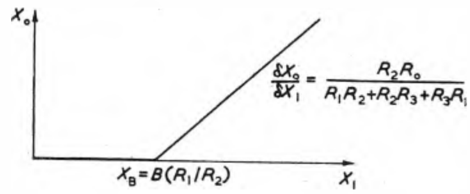


Fig. 15. Basic unit characteristic (Model A function generator)

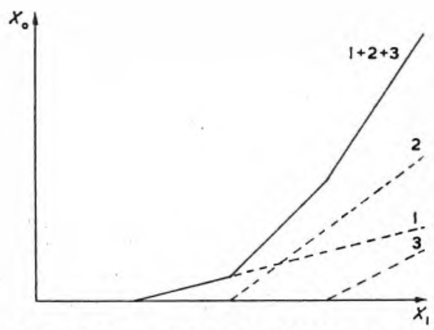


Fig. 16. Typical characteristic (Model A function generator)

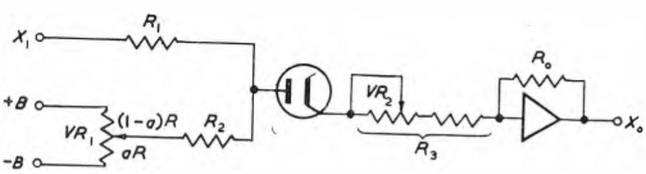


Fig. 17. General purpose arrangement (Model A function generator)

by break points of the individual units, and the slope is the sum of the slopes of the number of units with conducting diodes. Negative slopes can be obtained by switching the appropriate unit through a pre-amplifier to provide the required phase inversion.

For a general purpose flexible arrangement, potentiometer controls were used as in Fig. 17. The effective break point and slope are now given by:

$$X_B = B\frac{R_1(1-2a)}{R_2 + a(1-a)R}$$

$$\partial X_o/\partial X_1 = \frac{R_o[R_2 + a(1-a)R]}{R_1R_2 + R_2R_3 + R_3R_1 + a(1-a)\,R\,(R_1 + R_3)}$$

For fixed values of R_1, R_2 R_o, R and B, the break point depends only on the setting of potentiometer VR_1, whereas the slope depends on the settings of both VR_1 and VR_2. Thus, direct calibration of individual networks was regarded as too cumbersome, and was effected by calibration of a master unit. The individual units may then be nulled against the master, adjusting first break point and then slope.

Model A suffers from the limitations that the additional calibration unit is required, that the accuracy of required slope depends on the summation of a number of individual slopes and that available slope values are somewhat restricted (for component values chosen, between 0·25 and 2·5). All these difficulties are overcome by the use of model B, with which the remainder of this article is concerned.

Model B Function Generator

Model B uses the biasing system as described in Reference 11, and its main advantage is that each break point and slope may be set independently. It can be used up to the maximum signal frequencies (7° phase-shift at 200c/s) which will be encountered in the repetitive operation of LACE. The frequency response may be extended by switching out a low pass-filter, which is required only when use is made of the 'rounding off' facility.

For a given function, calculation of the optimum values of segment slopes and break points involves a few hours desk work. The actual setting of the approximate 'linear' function takes a quarter of an hour with a resultant accuracy of 1 per cent of full scale, that is, the output voltage representing the 'linear' function will depart from its desired value by not more than 0·5V.

Setting can be done either by visually matching the function generator characteristic, displayed on an oscilloscope, to the required function, drawn to scale on a graticule, or by accurate fixing of each break point and slope by d.c. null methods. For the first method, a sawtooth waveform is used to sweep the abscissa of the function and to provide an external time-base for the oscilloscope. A low-sweep speed is available for pen records.

The function generator provides a maximum of seven break points and eight slopes. Break points and slopes are continuously variable and the maximum value of slope is 7·5 (corresponding to 82°).

The facility of 'rounding off' any break point is provided, and is effected by modulation of the break point with a 20kc/s triangular waveform.

BASIC ARRANGEMENT

The basic arrangement is shown in Fig. 18, which, for simplicity, considers only three break points. B_1, B_2, B_3 are low impedance d.c. sources (to which may be added the 20kc/s triangular wave). The d.c. supplies are adjusted so that $B_3 > B_2 > B_1$. The diodes switch the inputs to the slope cathode-followers and these inputs may be either a constant break point voltage or the input voltage X and depend on the value of X.

The switching for all values of X is as follows:

	INPUT CF_1	INPUT CF_2	INPUT CF_3	INPUT CF_4
$X<B_1$	X	B_1	B_2	B_3
$B_1<X<B_2$	B_1	X	B_2	B_3
$B_2<X<B_3$	B_1	B_2	X	B_3
$B_3<X$	B_1	B_2	B_3	X

Thus, for the particular case when $B_1 < X < B_2$

$$F(X) = G(K_1B_1 + K_2X + K_3B_2 + K_4B_3)$$

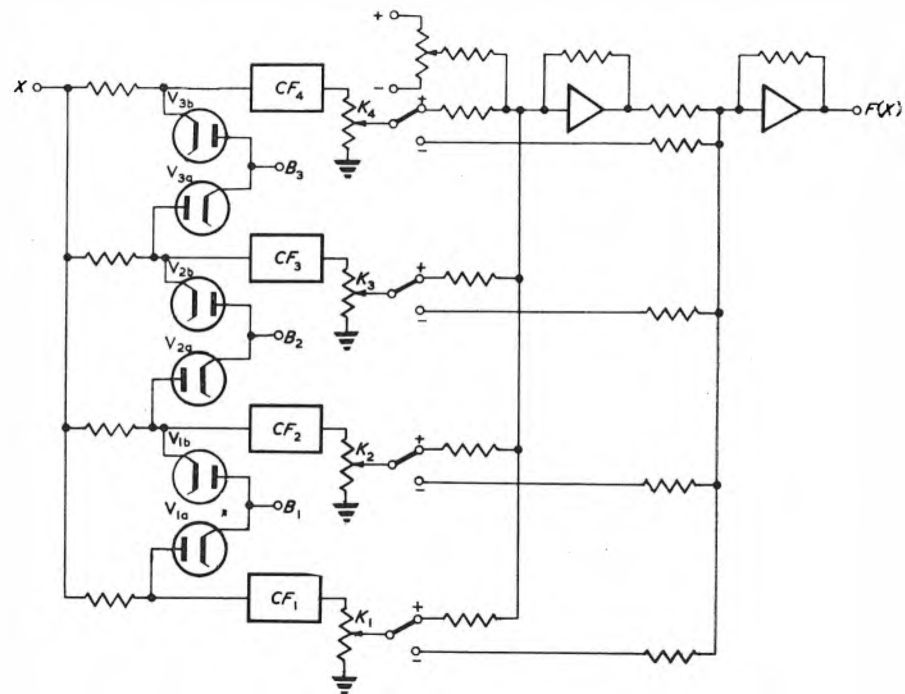
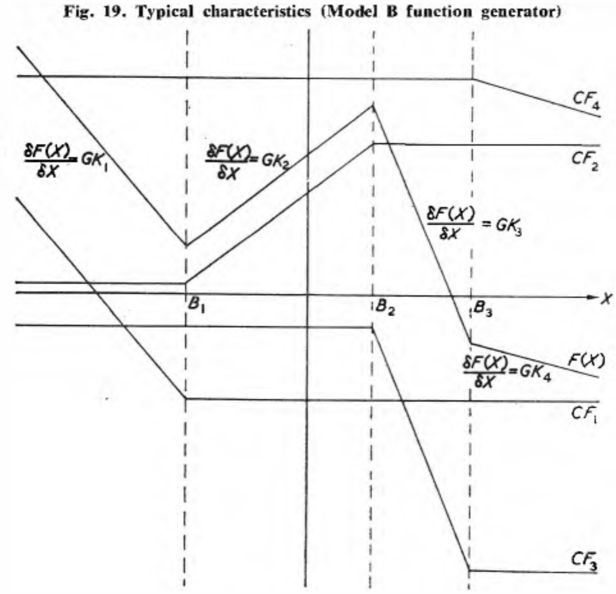


Fig. 18. Basic arrangement (Model B function generator)

Fig. 19. Typical characteristics (Model B function generator)



and $\delta F(X)/\delta X) = GK_2$

where G is the overall gain of the two d.c. feedback amplifiers, and for $G = 10$ this gives a maximum available slope of the same value.

Thus, using this arrangement, a characteristic of the type shown in Fig. 19 may be generated. Negative slopes are obtained by switching the appropriate cathode-follower direct to the input of the second amplifier, so that X experiences one instead of two phase inversions. It is to be noted that each break point and slope is controlled by independent adjustments.

In the case of Fig. 20, $F(X)$ is given by

$$G(-K_1B_1 + K_2X - K_3B_2 - K_4B_3) \text{ for } B_1 < X < B_2.$$

For $X = 0$, $F(0) = -G(K_1B_1 + K_3B_2 + K_4B_3)$. Having set in the required break points and slopes, the absolute value of the function for zero input depends on all the break point values and all slope values excepting the one operative through $X = 0$. In order to obtain the required $F(0)$, that is, the correct overall level of the function, it is necessary to provide a level control as shown in Fig. 18. It is important that aged diodes with small contact p.d. be used. For a small input voltage range about the break point, both diodes of a pair will conduct, so that, instead of a sharp transition between slopes, a zero slope region is interposed of width $\Delta_1 + \Delta_2$ where Δ_1, Δ_2 are the contact p.d.'s of the diode pair.

Also, the effective break point is modified if the break point source impedance is an appreciable fraction of the signal input resistance, and cathode-followers are therefore used to provide the voltage applied to the diode junctions.

'Rounding-off'

Any two adjacent linear segments may be replaced by a parabolic arc about the break point. If the voltage B is modulated by a high frequency triangular wave of amplitude A, the parabolic arc extends from $X = B - A$ to $X = B + A$ and will be tangential to the linear segments at these points. (The multiplier described above is a particular case of 'rounding off' when $A = B$ so that the square law characteristic applies from $X = 0$ to $X = 2A$.) This facility effects a considerable economy in the number of segments required to represent large changes of slope, as, for example, in the case of crests and troughs of a sinusoidal function.

The general expression for the parabola is

$$F(X) + \frac{1}{S_{n+1} - S_n}\left[AS_{n+1}S_n - B_n(S_{n+1}^2 - S_n^2)\right] = \frac{S_{n+1} - S_n}{4A}\left[X + \frac{S_{n+1} + S_n}{S_{n+1} - S_n}A - B_n\right]^2$$

for a break point B_n with associated slopes S_n, S_{n+1}.

Description of Complete Unit

A functional block diagram of the complete unit is shown in Fig. 20. This comprises seven diode pairs giving seven break points and eight slopes. An input buffer amplifier A_1 is needed to provide a high input impedance. The summing amplifiers A_2 and A_3 are identical with the summing amplifier as used in the multiplier. The purpose of the slope cathode-followers is to present a high constant impedance at the input resistance-diode junctions. A low-pass filter is required following A_3 to remove 20kc/s and harmonics, yet maintaining a good frequency response. A three-section tapered ladder filter is used with corner frequency at 5kc/s. Low output impedance for the complete function generator is provided by means of an output cathode-follower. The biasing arrangement for this cathode-follower involves a signal attenuation of 0·75 and restricts the maximum available slope to 7·5.

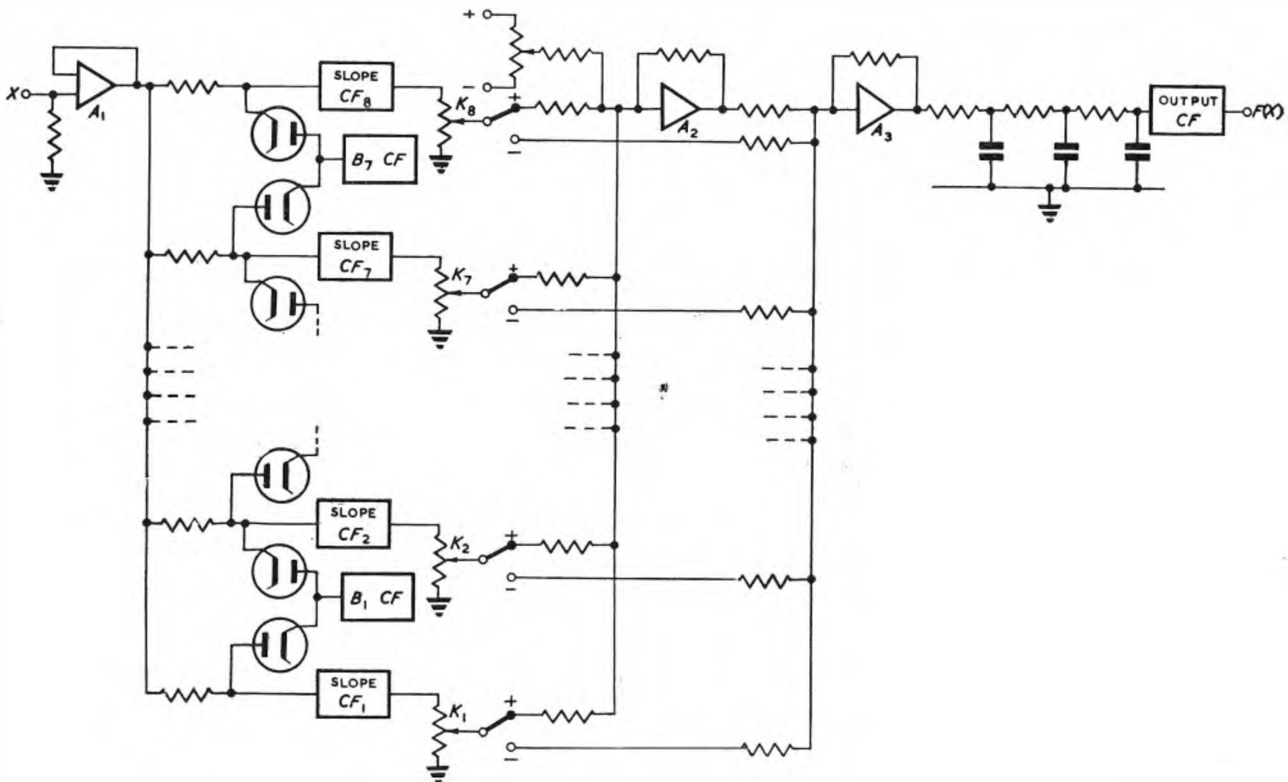


Fig. 20. Block diagram (Model B function generator)

Function Setting

This may be accomplished in two ways as indicated above. In method (a), the required function is drawn to a suitable scale on an oscilloscope graticule. A sawtooth waveform is applied to the X plates of the oscilloscope and to the input of the function generator. The output of the function generator is applied to the Y plates of the oscilloscope, and the break point and slope controls adjusted until the oscilloscope trace reproduces the function drawn on the graticule.

In method (b), to set slopes, the break points are adjusted so that the appropriate slope is operative through $X = 0$. The level control is then adjusted for $F(0) = 0$ and the output is then set by means of the slope control to its required value when a calibrated input voltage is applied. The outputs of the break point cathode-followers are adjusted to their calculated values by selecting each output in turn and balancing against a calibrated reference voltage divider. When all break point and slope adjustments have been made, the level control is adjusted to give the required value of the function for zero input.

Both of the above methods of function setting take about quarter of an hour.

A photograph of the front panel is shown in Fig. 21.

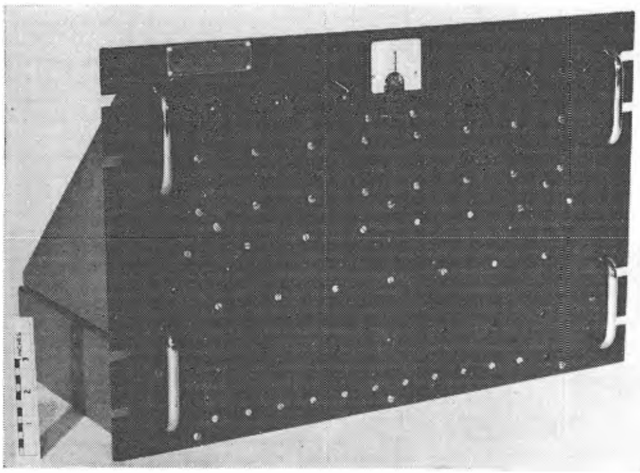


Fig. 21. Front panel view (Model B function generator)

Conclusions

The function generator described has the properties of flexibility and rapidity of function setting. The accuracy of reproduction of the approximate 'linear' function is 1 per cent of full scale. There is a limitation on accuracy due to the fact that the 'linear' function is made up of a maximum of eight slopes. This is most serious for a function with large rates of change of slope, but the limitation is offset by the provision of the 'rounding off' facility. An improvement in frequency response would be obtained by use of the LC low-pass filter, as described above for the multiplier.

Acknowledgments

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Appendix I

From Fig. 10

$$V_o[(1/R_1)+(1/R_2)+(1/R_3)+(1/R_4)] = (X/R_1)+(B/R_2)+Ap(t)/R_3 \quad \text{r.h.s.} < 0$$
$$= 0 \quad \text{r.h.s.} \geqslant 0$$

$$\text{i.e.} \quad \left.\begin{array}{ll} KV_o = [v - p(t)] & \text{r.h.s.} < 0 \\ = 0 & \text{r.h.s.} > 0 \end{array}\right\} \dots\dots\dots\dots (1)$$

where $K = (R_3/AR_E)$

$$(1/R_E) = (1/R_1)+(1/R_2)+(1/R_3)+(1/R_4)$$
$$v = [(X/R_1) + (B/R_2)]\ R_3/A$$

$p(t)$ is a periodic triangular wave (not necessarily symmetrical) of period T and in the interval $(0, T)$ one can write

$$\left.\begin{array}{ll} p(t) = (2t/m) - 1 & 0 < t < m \\ = \dfrac{2t}{T-m} + \dfrac{T+m}{T-m} & m < t < T \end{array}\right\} \dots\dots\dots (2)$$

Substituting equation (2) in equation (1)

$$\left.\begin{array}{ll} KV_o = (v - 1 + (2t/m) & \text{r.h.s.} < 0 \\ = 0 & \text{r.h.s.} > 0 \end{array}\right\} 0 \leqslant t \leqslant m$$

$$\left.\begin{array}{ll} = \left(v + \dfrac{T+m}{T-m} - \dfrac{2t}{T-m}\right) & \text{r.h.s. } 0 \\ = 0 & \text{r.h.s. } 0 \end{array}\right\} m \leqslant t \leqslant T$$

But

$$v - 1 + (2t/m) < 0 \quad \text{for } t < (1 - v)(m/2) = T_1$$

and

$$\left(v + \frac{T+m}{T-m} - \frac{2t}{T-m}\right) < 0 \text{ for } t > \left(v + \frac{T-m}{T-m}\right)\frac{T-m}{2} = T_2$$

Since $V_o(t)$ is periodic insofar as v can be regarded as a constant containing as it does frequencies much less than those of $p(t)$ one can write

$$V_o = a_o + \sum_n a_n \sin n\omega t - b_n \cos n\omega t$$

where $a_o = (1/T) \int_0^T V_o dt$

a_o representing in this case the d.c. content of V_o.

Hence

$$(1/K)\,a_o = (1/T)\int_0^{T_1} (v - 1 + (2t/m))\,dt - (1/T)\int_{T_2}^{T} v + \frac{T+m}{T-m} - \frac{2t}{T-m}\,dt$$

$$= (v-1)(T_1/T) + (T_1^2/T_m) + v + \frac{T+m}{T-m} - \frac{T}{T-m} - \left(v + \frac{T+m}{T-m}\right)(T_2/T) + \frac{T_2^2}{T(T-m)}$$

Substituting T_1 and T_2

$(1/K)\,a_o =$

$$\frac{-(1-v)^2 m}{2T} + \frac{(1-v)^2 m}{4T} + v + \frac{m}{T-m} - \left(v + \frac{T+m}{T-m}\right)^2 \frac{T-m}{2T} + \left(v + \frac{T+m}{T-m}\right)^2 \frac{T-m}{4T}$$

$$= \left(v + \frac{m}{T-m}\right) - \frac{(1-v)^2}{4T} - \left(v + \frac{T+m}{T-m}\right)^2 \frac{T-m}{4T}$$

$$= \frac{-(1-v)^2}{4}$$

i.e. $V_{o(dc)} = -(AR_E/4R_3)\,[(X/R_1 + B/R_2)\,R_3/A - 1]^2$

Thus if $(BR_3/AR_2) = 1$

$$V_{o(dc)} = -(R_E R_3/4R_1 A)X^2 \text{ for } 0 \geqslant X \geqslant -2A$$

Sim—If now the diode is reversed and $(BR_3/AR_2 = -1$

$$V_{o(dc)} = +(R_E R_3/4R_1 A)\,X^2 \text{ for } 0 \leqslant X \leqslant 2A.$$

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Some Characteristics of Metallic Varistors

By G. W. Holbrook*, Ph.D., A.M.I.E.E. and A. L. Dulmage*, M.A., Ph.D.

The application of metallic rectifiers in roles other than that of conventional rectification has led to an increased interest in the behaviour of these devices over the range of voltage which exhibits the maximum variation of resistance. In this article, an attempt has been made to provide an equivalent circuit which is applicable to all values of voltages, both positive and negative, within the rated limits of a given rectifier. From this circuit an expression has been established for describing the performance of the rectifier when used as a variable resistor or varistor.

THE term varistor is now commonly used to describe passive non-linear resistive circuit elements. This term embraces the complete field of two terminal passive resistors which exhibit non-ohmic or voltage dependent resistance characteristics and includes all types of barrier layer and point contact rectifiers, symmetrical non-linear resistors such as Thyrite and also the conventional thermionic diode. In general, varistors exhibit a practical resistance variation of several decades over their normal rated working range of current or voltage. In addition they usually possess an inherent capacitance which although very low in some cases, is generally variable with both applied voltage and frequency.

The object of this article is to outline some methods of dealing with this feature of variable resistance for varistors composed of point contact or barrier layer rectifiers. Practical equivalent circuits are suggested which represent these devices over the whole of their rated working range and formulæ are established for calculating the action of these elements when used in conjunction with linear resistive elements. No discussion of the effects of inherent self-capacitance is included, neither is any consideration given to the effect of temperature changes.

Varistor Characteristics

Because of the wide range of resistances existing within rated voltage limits it has become customary to illustrate the properties of rectifiers by plotting performance curves by either of the following methods:

(a) *Linear Plot of Voltage and Current*

This method is normally employed for displaying the characteristics of power type metal rectifiers and the current scale in the non-conducting direction is frequently made smaller than that of the conducting direction by one or more powers of ten. With this type of display the corresponding values of current and voltage define the static or d.c. resistance of the varistor. The incremental or dynamic resistance of the varistor to small signals can be defined as the inverse of the slope of this characteristic at any point. This method of presentation does not, however, allow an easy assessment of the change of d.c. resistance.

(b) *Semi-Logarithmic Plot of Resistance and Voltage*

This is the more usual method of displaying the characteristics of small barrier-layer and point contact rectifiers and is illustrated in Fig. 1. It is evident that the samples depicted exhibit an exponential relation between voltage and resistance over a considerable portion of the rated voltage range. Because of the comparative ease of manipulation of exponential functions this concept has been used throughout this article as a basis for discussing the action of all types of rectifier varistors.

The rating of a varistor depends upon its permissible power dissipation. For point contact and barrier layer rectifiers this is usually quoted in terms of the maximum permissible forward current and the maximum permissible reverse voltage that can be applied. The equivalent circuits established below are applicable only within the rated range of the rectifiers in question.

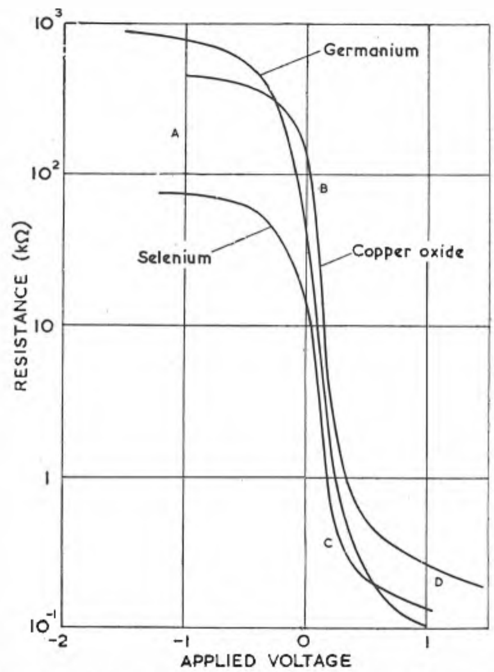

Fig. 1. Typical resistance characteristics of metallic varistors

Equivalent Circuits

Metallic rectifiers form a large group of varistors and embrace a wide range of materials, the more commonly used combinations being:

(a) Copper-copper oxide
(b) Selenium-iron
(c) Germanium-indium

Fig. 1 illustrates the resistance characteristics of a sample of each of these types. It will be noted that there are three main regions in each characteristic.

A-B—Where the applied voltage is negative and the resistance is high and reasonably constant.

B-C—At low values of positive voltage where an approximately exponential relation exists between resistance and voltage.

C-D—At higher values of positive voltage where the resistance is low and again comparatively constant in value.

For many applications, such as the process of conventional rectification, it is sufficiently accurate to ignore the second of these regions. Under this condition, the rectifier may be looked upon as a switching device which alternates

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the resistance of the circuit between the maximum and minimum values indicated at A and D respectively. For many other applications, the rectifier must, however, be looked upon as a continuously variable resistor and as such a more detailed consideration of its resistive properties must be made.

Beranek[1] has suggested that the relation between current and voltage can be expressed, as an approximation, for single plate copper-oxide rectifiers as follows:

$$I = k_1V + k_2V^2 \text{ for } 0 \leqslant V \leqslant 0{\cdot}02 \quad \ldots\ldots\ldots (1)$$
$$= k_3(QV - 1) \text{ for } 0{\cdot}06 \leqslant V \leqslant 0{\cdot}25 \quad \ldots\ldots\ldots (2)$$
$$= k_4V \quad \text{for} \quad V > 0{\cdot}25 \quad \ldots\ldots\ldots (3)$$

where k_1 to k_4 and Q are constants peculiar to any given rectifier. These relations are at best only approximate and

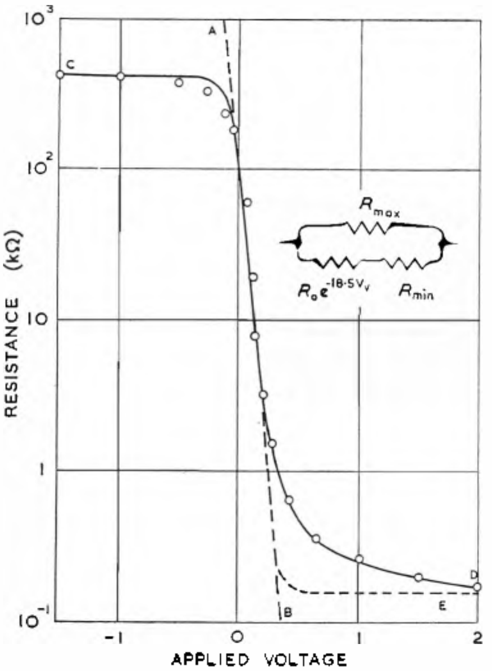

Fig. 2. Equivalent circuit for a copper oxide varistor

are limited to small regions of the characteristic. Tucker[4] has shown that a more accurate representation, true for a variety of varistors, can be expressed in the relation:

$$R = R_0\,e^{-gV} + R_{min} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

where R_{min} is a low and constant value of resistance obtained by a rectifier at the extreme limit of its forward current rating. R_0 in this equation is the intercept of the straight line portion of the resistance characteristic extrapolated back to the zero voltage axis, and g represents the slope of this straight line region. This concept holds reasonably well over the greater part of the positive voltage region but, of course, takes no account of the flattening of the resistance characteristic for negative voltages.

A more accurate equivalent circuit can be set up as shown in Fig. 2 in which the marked points represent the measured values of resistance of a copper-copper oxide rectifier. The line AB represents the extrapolation of the exponential portion of the characteristic. An estimate of the ultimate minimum resistance is 150Ω and the curve AE represents the characteristic as given by the approximation in equation (4). If, however, the equivalent circuit of the rectifier is assumed to be as shown in the figure then the synthetic characteristic is given by the curve AD.

This equivalent circuit is similar to that suggested by Lovelock[3] for a point contact germanium rectifier. It differs from the concept of equation (4) in that the value of the voltage in the exponent is not the measured voltage across the rectifier, V_w, but a postulated voltage V_v which exists across a variable resistor R_v such that:

$$R_v = R_0\,e^{-gV_v} \quad \ldots\ldots\ldots\ldots\ldots\ldots (5)$$

It also differs from the classical concept which defines the current flowing across the junction of a *pn* rectifier as:

$$I = I_0(e^{QV} - 1) \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

where I_0 is a constant, V the applied voltage and Q a further constant peculiar to the material of the junction. This equation ignores the bulk resistance of the junction and thus corresponds with equation (5) for the value of R_v. From equation (6) the resistance of the *pn* junction for very small values of voltage can be written as:

$$R_v = \lim_{V \to 0} (dV/dI) = \lim_{V \to 0} (1/QI_0)\,e^{-QV}$$

Thus the factor $1/QI_0$ is the equivalent of R_0 in equation (5), but Q is not equal to g and equation (5) can only be considered as empirical from the outset.

The addition of the ohmic resistance R_{min}, representing the bulk resistance of the rectifier, leads to the following approximation for the forward resistance of a rectifier:

$$R_w = R_0e^{-gV_v} + R_{min} \quad \ldots\ldots\ldots\ldots\ldots (7)$$

A further addition of a shunt resistor R_{max} to the rectifier leads to the final equivalent circuit as shown in Fig. 2. The value of this resistance in the figure is chosen as 440kΩ, giving the reverse voltage characteristic as shown in the line CD. The final equation for the resistance of the rectifier for all values of voltages now becomes:

$$R_w = \frac{R_{max}(R_0e^{-gV_v} + R_{min})}{R_{max} + R_0e^{-gV_v} + R_{min}} \quad \ldots\ldots\ldots (8)$$

Since R_{max} is always much greater than R_{min} an alternative arrangement is possible in which R_{max} is in shunt with R_v and R_{min} is placed in series with this combination. The equation for this arrangement becomes:

$$R_w = \frac{R_{max}R_0e^{-gV_v}}{R_{max} + R_0e^{-gV_v}} + R_{min} \quad \ldots\ldots\ldots\ldots (9)$$

For all practical rectifiers there is no significant difference between these two alternative equivalent circuits.

Standardization of Samples

One of the practical limitations in the employment of metal rectifiers in precision circuits is the wide range of characteristics obtained in a given batch of samples. This is illustrated in Fig. 3 which shows the extreme variation of characteristics, curves AB and CD, obtained within a batch of ten similar copper oxide rectifiers. These characteristics show a wide variation in the values of R_{min} and R_{max}. A method, similar to that suggested by Harvey[2] for correcting deviations in thermionic valve characteristics, and consisting of adding series (padding) and parallel (trimming) resistors can be applied to minimize these deviations. To accomplish this a standard value of R_{min}, greater than the maximum observed, and a standard value of R_{max} smaller than the minimum observed must be chosen. In Fig. 4 standard values were chosen as follows:

$R_{max} = 1\text{M}\Omega$
$R_{min} = 700\Omega$

The original extreme samples, when trimmed and padded with external series and parallel resistors as shown, result in the standard characteristic depicted at EF. This method of standardization suggests that individual samples, although exhibiting major discrepancies in R_{max} and R_{min} do maintain a reasonably constant value of g, and that, in practice, one is not limited to a method of selecting samples in order to maintain uniformity. The main disadvantage lies in the overall lessening of the resistance range of the standardized samples.

Symmetrical Varistors

It is evident that if two standardized rectifiers are connected in parallel but in opposite mutual polarity then a symmetrical varistor is formed which will produce identical current-voltage relations for either direction of current flow. Such an arrangement can be represented by one of two alternative equivalent circuits. In the first of these, both values of R_{max} and R_v are assumed to be in parallel with each other and the whole in series with a single resistor of R_{min}. This arrangement leads to an equation for the total

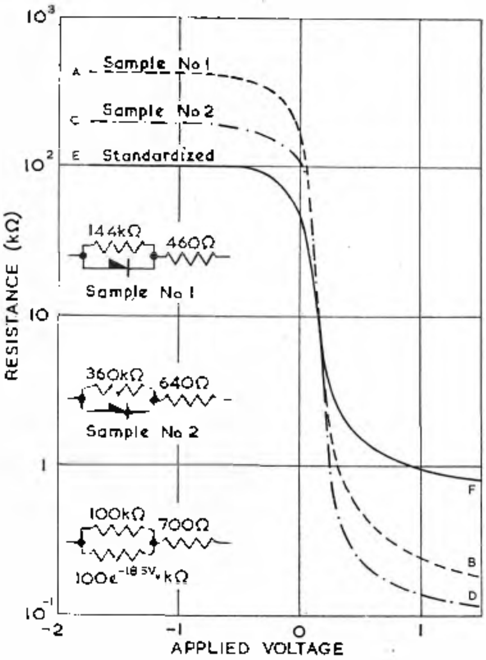

Fig. 3. Standardization of a batch of copper oxide varistors

resistance of the symmetrical varistor of:

$$R_w = \frac{R_{max}}{2\left[\frac{R_{max}\cosh gV_v}{R_0} + 1\right]} + R_{min} \quad \ldots\ldots\ldots\ldots (10)$$

The second alternative is represented by the two values of R_v, in parallel, with a single resistor R_{min} in series with this combination. The two values of R_{max} are then assumed to be in parallel with the whole, giving:

$$R_w = \frac{R_{max}[R_0 + 2R_{min}\cosh gV_v]}{2[R_0 + 2(R_{min} + (R_{max}/2))\cosh gV_v]} \quad \ldots\ldots (11)$$

It is evident from both equations (10) and (11) that the maximum resistance of the varistor, achieved as V_v approaches zero, is given by the approximation:

$$R_w \simeq \frac{1}{2}\left[\frac{R_0R_{max}}{R_0 + R_{max}}\right] \quad \ldots\ldots\ldots\ldots (12)$$

and that the minimum resistance, for large values again approaches the value R_{min}.

Multiple Varistors

If two or more similar varistors are connected in series and in the same polarity then a composite varistor is formed whose resistance for n like sections is given by:

$n^2R_{max}R_0 \exp(-gV_v/n)$

$$R_w = \frac{n^2R_{max}R_0\exp(-gV_v/n)}{n(R_{max} + R_0\exp(-gV_v/n)} + nR_{min} \quad \ldots\ldots (13)$$

where R_{max}, R_{min} and R_0 are the constants of one section and V_v is now the voltage developed across the composite variable resistor.

For n similar varistors connected in parallel the corresponding expression is:

$$R_w = (1/n)\left[\frac{R_{max} + R_0e^{-gV_v}}{R_{max}R_0e^{-gV_v}}\right] + (R_{min}/n) \quad \ldots\ldots (14)$$

The equivalent circuits established for both symmetrical and non-symmetrical varistors include one or more variable resistors R_v whose value depends upon the magnitude and direction of a voltage V_v developed across this element. The equivalent circuit also includes a fixed resistor R_{min} representing the theoretical minimum resistance of the varistor. In practical circuits other external resistors may

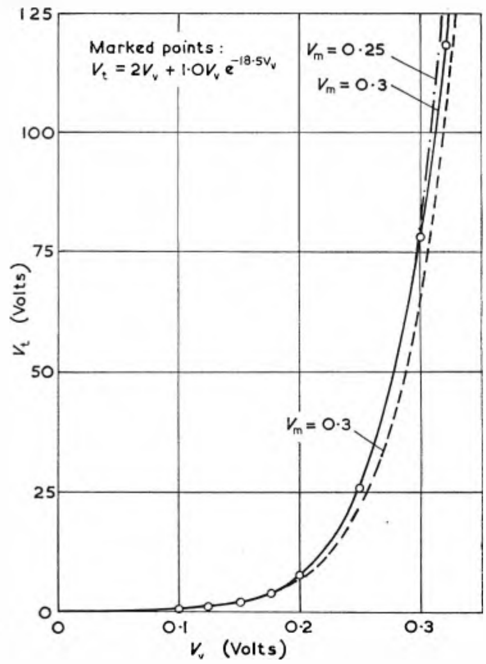

Fig. 4. Illustrating the accuracy of the relation $V = C\ \mathrm{Sinh}\ DV_v$

also be added in series with the varistor. In order to establish the ratio of the voltage developed across the variable resistor, V_v, to total voltage applied to the whole circuit, V_t, it is evident that some equation must be established which defines V_v in terms of either V_w or V_t.

Non-Symmetrical Varistors

From equation (8) it can be shown that the value of V_w, the voltage developed across the complete varistor, can be written in terms of V_v as:

$$V_w = \left[\frac{R_vR_{max}/(R_v + R_{max}) + R_{min}}{R_vR_{max}/(R_v + R_{max})}\right] V_v$$

$$= (1 + (R_{min}/R_{max}))\, V_v + (R_{min}/R_0)V_ve^{gV_v}$$

$$= AV_v + BV_v e^{gV_v} \quad \ldots\ldots\ldots\ldots\ldots\ldots (15)$$

where A and B depend upon the parameters of the equivalent circuit. If a further resistor R is added in series with the varistor then the total voltage developed across the whole circuit can now be expressed as:

$$V_t = \left(1 + \frac{R_{min} + R}{R}\right)V_v + \frac{R_{min} + R}{R_0}V_ve^{gV_v} \quad \ldots (16)$$

which is similar to equation (15) but with different values of A and B. Both of these equations are difficult to express in inverse form due to the added term AV_v. However, the whole right-hand side of each equation can be made to approximate to a hyperbolic sine function, as shown below.

Assume that the following expression holds for positive values of V_v:

$$V = C\sinh DV_v \quad \ldots\ldots\ldots\ldots\ldots (17)$$

A theoretical equivalence of equations (16) and (17) can be established at three points, viz:

(1) $V_v = 0$ This condition is already assumed in equation (17).

(2) $V = V_m$ Where V_m is the maximum value of V_v attained at the maximum forward current of the varistor.

(3) $V_v = (V_m/2)$

Taking conditions (2) and (3), the following equations can now be written:

$$C \sinh V_m D = AV_m + BV_m e^{gV_m} \quad \ldots\ldots\ldots\ldots (18)$$

$$C \sinh (V_m D/2) = (AV_m/2) + (BV_m/2)\, e^{gV_m/2} \quad \ldots (19)$$

Dividing equation (18) by equation (19) gives:

$$\frac{\sinh V_m D}{\sinh (V_m D/2)} = f(A, B, V \text{ and } g) = L \ldots\ldots\ldots (20)$$

whence:

$$e^{V_m D} - e^{-V_m D} = L\,[e^{V_m D/2} - e^{-V_m D/2}]$$

Now let $e^{V_m D/2} = M$ then:

$$M^2 - LM = M^{-2} - LM^{-1}$$

Completing squares and taking the negative alternative gives:

$$M^2 - LM + 1 = 0$$

Solving for M then gives:

$$e^{V_m D/2} = M = \frac{L \pm \sqrt{(L^2 - 4)}}{2}$$

and

$$D = (2/V_m) \operatorname{Ln}\left[\frac{L + \sqrt{(L^2 - 4)}}{2}\right] \ldots\ldots (21)$$

and

$$C = \frac{AV_m + BV_m e^{gV_m}}{\sinh DV_m} \ldots\ldots\ldots\ldots (22)$$

These last two equations are general in form and in terms of the measurable parameters established in the corresponding equivalent circuit. The accuracy with which they represent the original exponential equation established in equation (16) depends upon the range of voltage covered by the value of V_m. This is illustrated in Fig. 4 in which the corresponding values of V_t and V_v have been plotted for a copper oxide varistor (Sample No. 5) where $R = R_{min} \simeq R_{max} \approx R_o$. The marked points show the calculated relation:

$$V_t = 2V_v + V_v e^{18 \cdot 5 V_v}$$

and the curves represent the values obtained as:

$V_t = 0{\cdot}183 \sinh 22{\cdot}5 V_v$ for $V_m = 0{\cdot}3$V

$V_t = 0{\cdot}179 \sinh 22{\cdot}7 V_v$ for $V_m = 0{\cdot}25$V

$V_t = 0{\cdot}235 \sinh 21{\cdot}2 V_v$ for $V_m = 0{\cdot}35$V

The choice of V_m must obviously be made so that it represents the true working range of V_v for the particular case in question.

The foregoing results (equations (17 to 22)) are valid for positive or forward voltages only. For reverse voltages, a reasonable approximation results if the backward resistance of the varistor is assumed to be constant and equal to R_{max}. Under these conditions, and for reverse voltages, the total voltage developed across a non-symmetrical varistor in series with a linear resistor can be written as:

$$V = \frac{V_v(R + R_{min} + R_{max})}{R_{max}} \ldots\ldots\ldots (23)$$

This approximation will show significant errors only in the region where V_t is small and when R_o is considerably smaller than R_{max}.

Symmetrical Varistors

The relation between V_w and V_v for a symmetrical varistor can be written directly from equation (10) as:

$$V_w = V_v[1 + (2R_{min}/R_{max})] + V_v(2R_{min}/R_o) \cosh gV_v$$

or from equation (11) as:

$$V_w = V_v + V_v(2R/R_o) \cosh PV_v$$

If a combination of a symmetrical varistor and a linear resistor, R is considered the following equations apply:

$$V_t = V_v\left[1 + \frac{2(R_{min} + R)}{R_{max}}\right] + V_v \frac{2(R_{min} + R)}{R_o} \cosh gV_v \ldots\ldots\ldots (24)$$

$$V_t = V_v + V_v \frac{2(R_{min} + R)}{R_o} \cosh gV_v \ldots\ldots (25)$$

In either case the relation can be written in general terms as:

$$V_t = EV_v + FV_v \cosh gV_v \ldots\ldots\ldots\ldots (26)$$

The relations established in equations (21) and (22) can be applied directly to symmetrical varistors if equation (20) is written in the form:

$$L = \frac{\sinh V_m D}{\sinh (V_m D/2)} = \frac{EV_m + FM_m \cosh gV_m}{(EV_m/2) + (FV_m/2) \cosh (gV_m/2)} \ldots\ldots (27)$$

This again allows the synthetic equation of:

$$V_t = C \sinh DV_v$$

to be written in terms of the known parameters, E, F, and g. This last relation is, of course, applicable to all values of V_v both positive and negative. Thus, for non-symmetrical varistors, in the conducting condition and for symmetrical varistors for all values of voltage the explicit relation between V_v and V_t or V_w can be written in the form:

$$V_v = (1/D) \sinh^{-1} V_t/C \ldots\ldots\ldots\ldots (28)$$

For non-symmetrical varistors in the non-conducting condition the appropriate expression approximates to:

$$V_v = V_t \frac{R_{max}}{R + R_{min} + R_{max}} \ldots\ldots\ldots (29)$$

Incremental Impedance

The action of a rectifier under small-signal conditions is essentially that of a variable resistive impedance and, if the inherent self-capacitance of the varistor can be ignored, then the impedance can be written as:

$$Z_w = (dV_w/dI) \ldots\ldots\ldots\ldots\ldots (30)$$

where I is the current flowing through the varistor and V_w is the total bias voltage across it. This definition assumes that the incremental voltage, dV_w, is small compared with the total bias voltage and that the relationship between I and V_w is linear over this small voltage change.

It has been suggested by Tucker[2] that an approximate relation between the current and voltage in a varistor can be written as:

$$I = m\, e^{QV_w} \ldots\ldots\ldots\ldots\ldots\ldots (31)$$

where m is the intercept on the voltage axis made by extrapolating the exponential range of the current-voltage characteristic when plotted on a semi-logarithmic basis. This equation is only valid in the middle range of positive voltages as it assumes that a finite current, m, flows at zero voltage and, in addition, the exponent QV_w is based on the total voltage across the varistor rather than on V_v, developed across the variable resistor R_v. From this equation, however, an approximate expression can be obtained for the incremental impedance as:

$$Z_w = (1/mQ)e^{-QV_w} - R_{min} \ldots\ldots\ldots (32)$$

A more accurate expression for the impedance of a rectifier can be obtained by considering the equivalent circuit described in equation (8). The total current through the rectifier can now be written as:

$$I = \frac{V_v[R_{max} + R_o e^{-gV_v}]}{R_{max} R_o e^{-gV_v}} \ldots\ldots\ldots\ldots (33)$$

whence

$$dI/dV = \frac{e^{-gV_v}(1 - gV_v)R_{max} + R_o}{R_{max}R_o}$$

The total impedance of the rectifier can now be written as:

$$Z_w = (dV_v/dI) + R_{min} = \frac{R_{max} R_0}{R_0 + R_{max} e^{-gV_v}(1+gV_v)} + R_{min} \quad \ldots\ldots\ldots\ldots (34)$$

and V_v can be defined by the implicit relation given in equation (15). Equation (34) is valid for all values of V_w, both positive and negative, but suffers from the disadvantage that the unknown value of V_w must be calculated for each value of V_w.

An approximate expression for the impedance of a rectifier for forward values of bias, can be established directly in terms of V_w by using the relation:

$$V_w = C \sinh DV_v$$

The current flowing through the rectifier can be written as:

$$I = \frac{V_w - V_v}{R_{min}}$$

$$= (1/R_{min})\, V_w - (1/D) \sinh^{-1} (V_w/C)$$

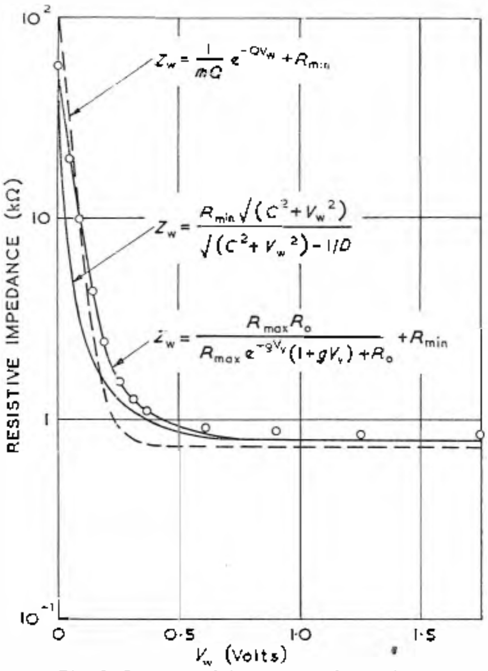

Fig. 5. Incremental impedance of a varistor

whence

$$Z_w = (dV_w/dI) = R_{min} \frac{\sqrt{(C^2 + V_w^2)}}{\sqrt{(C^2 + V_w^2)} - (1/D)} \quad \ldots (35)$$

A comparison of these methods is illustrated in Fig. 5 which shows the values of resistive impedance of the standardized oxide rectifier, illustrated in Fig. 3, measured at 1kc/s. The curves shown are based on the previously established equations (32), (34) and (35). This comparison indicates that equation (34) is by far the most accurate method of calculating the incremental impedance of such a rectifier. Equation (32) is reasonably valid in the middle range of voltages but its accuracy deteriorates at the higher and lower extremes. Equation (35) gives an accurate representation of the impedance at the extremes but loses accuracy in the middle range of voltages due to the approximate nature of the explicit relation between V_w and V_v assumed at the outset.

Voltage Division

A more general case in the employment of varistors occurs under large-signal conditions, where the effective resistance of the varistor is changed by virtue of the signal applied to it. A frequently occurring situation of this type is one in which a varistor is connected in series with a linear resistor to form what is essentially a voltage divider. If an input voltage, V_t, is applied to the complete circuit then an output voltage can be selected either across the varistor, V_w, or across the linear resistor V_r.

This situation can be written thus:

$$V_t = V_r + V_w = V_r + V_v + V_o$$

where V_o is the voltage developed across the resistor R_{min}.

The output of such a voltage divider can be expressed as either:

$$V_r = V_t - V_w = V_t - (V_v + V_o) = V_t - V_v - (V_r R_{min}/R)$$

$$= V_t\left(\frac{R}{R + R_{min}}\right) - \left(\frac{R}{R + R_{min}}\right)(1/D) \sinh^{-1} (V_t/C) \quad \ldots\ldots\ldots\ldots\ldots (36)$$

or:

$$V_w = V_t - V_r = V_t - (V_o R/R_{min}) = V_t - \frac{(V_w - V_v)R}{R_{min}}$$

$$= V_t\left(\frac{R_{min}}{R + R_{min}}\right) + \left(\frac{R_{min}}{R + R_{min}}\right)(1/D) \sinh^{-1} (V_t/C) \quad \ldots\ldots\ldots\ldots\ldots\ldots (37)$$

These equations are valid for all values of voltage, both

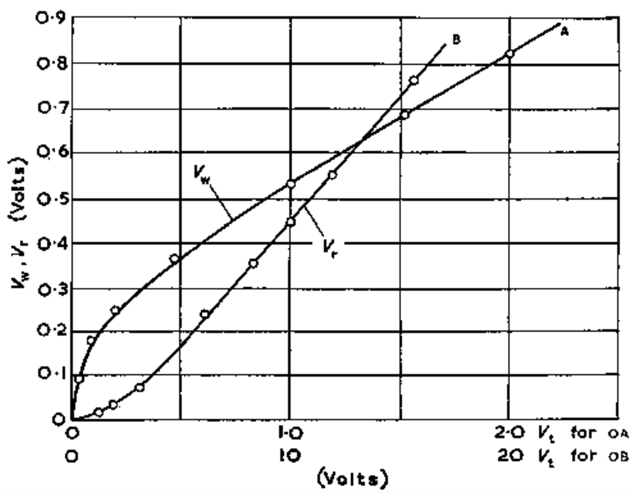

Fig. 6. Voltage division in a varistor and resistor combination

positive and negative, in the case of a symmetrical varistor. For a non-symmetrical varistor in the non-conducting condition the following approximations apply:

$$V_w = \left(\frac{R_{max}}{R + R_{max}}\right) V_t \quad \ldots\ldots\ldots\ldots (38)$$

$$V_r = \left(\frac{R}{R + R_{max}}\right) V_t \quad \ldots\ldots\ldots\ldots (39)$$

These forms of equations are quite adequate for calculating the relation between instantaneous values of V_t, V_w and V_r, as illustrated in Fig. 6. Curve OA illustrates the relation between V_w and V_t of a symmetrical varistor whose constants are:

$$R_{max} = 100\text{k}\Omega$$
$$R_{min} = 700\Omega$$
$$R_o = 100\text{k}\Omega$$
$$g = 18{\cdot}5$$

when connected in series with a resistor of 25kΩ. The relevant equation for this curve is given by:

$$V_w = 0{\cdot}027\, V_t + 0{\cdot}048 \sinh^{-1} 10V_t$$

and the marked points indicate measured values of V_w for this particular condition. The curve OB represents the respective values of V_r and V_t for the same varistor when connected in series with a resistor of 1kΩ. The calculated relation of this curve is:

$$V_r = 0{\cdot}06V_t - 0{\cdot}005 \sinh^{-1} 11{\cdot}8V_t$$

and the marked points again show the measured values.

Production of Harmonics

If the input signal to the voltage divider consists of a sinusoidal voltage which can be written as:

$$V_t = V_{max} \sin \theta = V_{max} \sin \omega t$$

then it is evident that harmonics of the fundamental frequency ω will be developed in both of V_r and V_ω. In the case of a symmetrical varistor these harmonics will consist of the odd orders only, but for a non-symmetrical varistor harmonics of all orders will be produced.

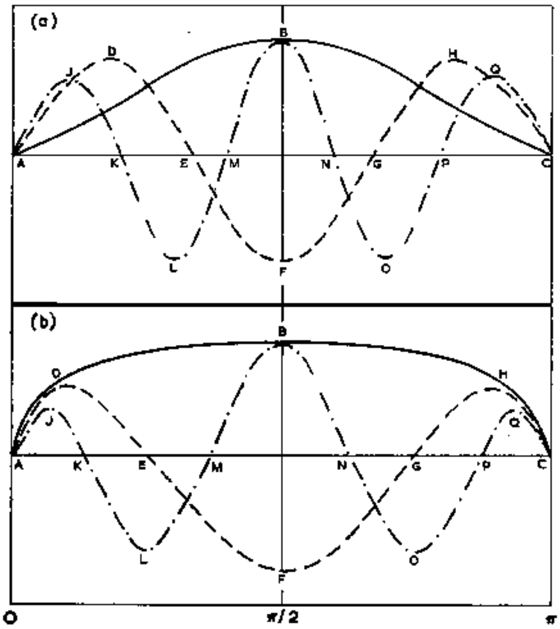

Fig. 7. Application of Simpson's Rule for integrating (Sinh^{-1} × sin θ) sin nθ

The general form of the output of a voltage divider containing a symmetrical varistor can be written as:

$$V_x = k_1 V_t + k_2 \sinh^{-1} k_3 V_t \quad \ldots\ldots\ldots\ldots (40)$$

where k_1, k_2, k_3 are given by the appropriate constants in equations (36) and (37).

This same form also applies to the output from a divider containing a non-symmetrical varistor, when in the conducting condition. Due to the divergent nature of the inverse hyperbolic function a direct calculation of the coefficients of harmonics cannot be made. However, a very close approximation adequate for practical purposes can be made as shown below.

(a) Symmetrical Voltage Dividers

If it is assumed that the general expression given in equation (40) can be written as:

$$V_x = k_1 V_{max} \sin \theta + b_1 \sin \theta + b_3 \sin 3\theta + b_5 \sin 5\theta \text{ etc.}$$

then the coefficients in this last expression can be determined by means of an expansion in a Fourier series. As V_t is an odd function of θ only the interval 0 to π need be considered and these coefficients can be written as:

$$b_1 = (2/\pi)\, k_2 \int_0^\pi \sinh^{-1} [X \sin \theta] \sin \theta \, d\theta$$

$$b_3 = (2/\pi)\, k_2 \int_0^\pi \sinh^{-1} [X \sin \theta] \sin 3\theta \, d\theta$$

$$b_5 = (2/\pi)\, k_2 \int_0^\pi \sinh^{-1} [X \sin \theta] \sin 5\theta \, d\theta$$

where $X = k_3 V_{max}$.

As there is no convenient method of performing these integrations, it is convenient to consider the graphs of the integrands and to apply Simpson's Rule to establish approximate average values of these functions. Fig. 7(a) shows a sketch of the corresponding values of:

(a) $\sinh^{-1} (X \sin \theta) \sin \theta$ at ABC
(b) $\sinh^{-1} (X \sin \theta) \sin 3\theta$ at ADEFGC
(c) $\sinh^{-1} (X \sin \theta) \sin 5\theta$ at AJKLBNOPQC

plotted against θ between the limits $\theta = 0$ and $\theta = \pi$. The coefficients b_1, b_3, b_5 are equal to twice the average ordinate for each harmonic and can be calculated as shown below:

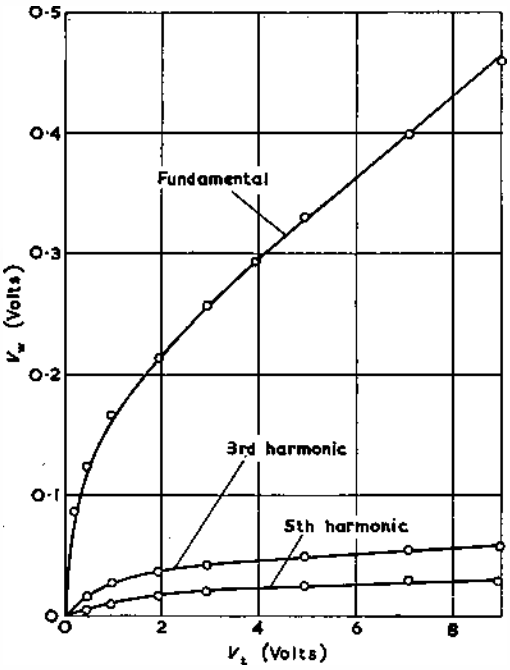

Fig. 8. Relative harmonics produced in a symmetrical varistor and resistor combination

(1) *Fundamental*

The curve drawn for the fundamental deviates from the form of a parabola and a more accurate application of Simpson's Rule is obtained by considering this curve in two halves:

Area under AB =

$$\pi/12 \; [0 + 4 \sinh^{-1} (X \sin \pi/4) \sin \pi/4 + \sinh^{-1} X]$$

The area under BC is identical and thus the total area is given by:

$$\pi/6 \; [4 \sinh^{-1} (X \sin \pi/4) \sin \pi/4 + \sinh^{-1} X]$$

The coefficient of the fundamental is given by:

$$k_1 V_{max} + b_1 = k_1 V_{max} - 4k_2/3 \, [\sinh^{-1} (X \sin \pi/4) \sin \pi/4 + (1/4) \sinh^{-1} X] \quad \ldots\ldots\ldots\ldots (41)$$

(2) *Third Harmonic*

The areas ADE, EFG and GHC are all approximately parabolic in form and hence by Simpson's Rule:

$$\text{Area ADE} = \text{Area GHC} = (2\pi/9) \sinh^{-1} \; (X \sin \pi/6 \sin 3\pi/6$$

$$\text{Area EFG} = (2\pi/9) \sinh^{-1} (X \sin \pi/2) \sin 3\pi/2$$

$$\text{Hence } b_3 = (4k_2/9) \, [2 \sinh^{-1} (X \sin \pi/6) \quad \sinh^{-1} X] \quad \ldots\ldots\ldots\ldots (42)$$

(3) *Fifth Harmonic*

$$b_5 = (2k_1/\pi) \, [\text{Area AJK} + \text{Area KLM} + \text{Area MBN} + \text{Area NOP} + \text{Area PQC}]$$

$$= (4k_2/15) \, [2 \sinh^{-1} (X \sin \pi/10) - 2 \sinh^{-1} (X \sin 3\pi/10) + \sinh^{-1} X] \quad \ldots\ldots\ldots\ldots (43)$$

(4) *General Term*

The coefficient of the n^{th} harmonic is zero if n is even.

If n is odd and greater than one then:

$$b_n = (4k_2/3n)\,[2\sinh^{-1}(X\sin\pi/2n) + \ldots$$
$$+ 2(-1)^r\sin^{-1}\left(X\sinh\frac{(2r+1)\pi}{2n}\right) + \ldots$$
$$+ 2(-1)^{(n-3)/2}\sinh^{-1}\left(X\sin\frac{(n-2)\pi}{2n}\right) + \ldots$$
$$+ (-1)^{(n-1)/2}\sinh^{-1}X] \quad \ldots\ldots\ldots (44)$$

The accuracy of this method can be judged from Fig. 8 which shows the calculated and measured values of the fundamental, the third and the fifth harmonics of V_w for a voltage divider. The solid curves represent the calculated values and the marked points represent measured values.

(b) Non-Symmetrical Voltage Dividers

Due to the asymmetry of the varistor in this case the expression for V_w or V_r in terms is no longer odd, but can be written in general terms as:

$$V_x = k_1V_{max}\sin\theta + k_2\sinh^{-1}(k_3V_{max}\sin\theta)$$

for $0 \leqslant \theta \leqslant \pi$ and:

$$V_x = k_4V_{max}\sin\theta$$

for $\pi \leqslant \theta \leqslant 2\pi$

One denotes the Fourier expression in this case as:

$$V_X = \alpha_0' + \alpha_1'\cos\theta + \alpha_2'\cos 2\theta + \ldots + \alpha_n'\cos n\theta + \ldots \text{etc.}$$
$$+ b_1'\sin\theta + b_2'\sin 2\theta + \ldots \quad + b_n'\sin n\theta + \ldots \text{etc.}$$

for $0 \leqslant \theta \leqslant 2\pi$

In such an expression if n is odd then $\alpha_n' = 0$ and consequently b_n' is the coefficient of the n^{th} harmonic. Conversely if n is even then $b_n' = 0$ and α_n' is the coefficient of the n^{th} harmonic. Fig. 7(b) shows the graphs of the functions:

$\sinh^{-1}(X\sin\theta)$ at ABC

$\sinh^{-1}(X\sin\theta)\cos 2\theta$ at ADEFGHC

$\sinh^{-1}(X\sin\theta)\cos 4\theta$ at AJKLMBNOPQC

where $X = k_3V_{max}$. These integrands occur in connexion with the even harmonics and correspond with those shown in Fig. 7(a) for the odd harmonics. The coefficients of the d.c. component and the even order harmonics are found from the average values as shown below:

(1) *D.C. Component*

α_0' = Mean value for V for $\theta \leqslant 0 \leqslant 2\pi$

$$= (1/2\pi)\left[\int_0^\pi [k_1V_{max}\sin\theta + k_2\sinh^{-1}X\sin\theta]d\theta + \int_\pi^{2\pi} k_4V_{max}\sin\theta\, d\theta\right]$$

$$= \frac{(k_1 - k_4)V_{max}}{\pi} + (k_2/2\pi)\,[\text{Area ABC}]$$

and using Simpson's Rule:

$$\alpha_0' = \frac{(k_1 - k_4)V_{max}}{\pi} + (k_2/3)\sinh^{-1}X \quad \ldots\ldots\ldots (45)$$

(2) *Second Harmonic*

$$\alpha_2' = \frac{2(k_4 - k_1)V_{max}}{3\pi} + (1/\pi)\,[\text{Area ADE} + \text{Area EFG} + \text{Area GHC}]$$
$$= \frac{2(k_4 - k_1)V_{max}}{3\pi} + (k_2/3)\,[\sinh^{-1}(X\sin\pi/8)\cos\pi/4 - \sinh^{-1}X] \quad \ldots\ldots\ldots (46)$$

(3) *Fourth Harmonic*

$$\alpha_4' = \frac{2(k_4 - k_1)V_{max}}{15\pi} + (1/\pi)\,[\text{Area AJK} + \text{Area KLM} + \text{Area MBN} + \text{Area NOP} + \text{Area PQC}]$$
$$= \frac{2(k_4 - k_1)V_{max}}{15\pi} + k_2/6[\sin^{-1}(X\sin\pi/16)\cos\pi/4 - 2\sinh^{-1}(X\sin\pi/4) + \sinh^{-1}X)] \quad \ldots\ldots\ldots (47)$$

(4) *General Even Harmonic*

$$\alpha_n' = \frac{2(k_4 - k_1)V_{max}}{\pi(n^2 - 1)} + (2k_2/3n\,[\sinh^{-1}(X\sin\pi/4n)\cos\pi/4 + 2\sum_{r=1}^{(n/2)-1}(-1)^r\sinh^{-1}(X\sin r\pi/n) + (-1)^{n/2}\sinh^{-1}X] \quad \ldots\ldots (48)$$

(5) *Odd Harmonics*

The coefficient of the fundamental is found by integrating $V_x\sin\theta$ over the whole range 0 to 2π and becomes:

$$b_1' = \frac{(k_1 + k_4)V_{max}}{2} + b_1/2 \quad \ldots\ldots\ldots (49)$$

where b_1 is coefficient of the fundamental as obtained in the case of a symmetrical varistor. Since:

$$\int_\pi^{2\pi} V_x\sin n\theta\, d\theta = 0$$

for n odd and greater than unity then coefficients of the third and higher odd order harmonics can be obtained by integrating $V_x\sin n\theta$ over the range 0 to π and averaging the result over the complete range 0 to 2π. This yields the following general result:

$$b_n' = (b_n/2) \quad \ldots\ldots\ldots (50)$$

An example of this method of computation is given in Table 1 which shows the calculated and measured values of the constituents of V_r for a non-symmetrical voltage divider. This voltage divider was exactly similar to that shown in Fig. 8 except that one section only of the symmetrical varistor was employed.

TABLE 1.
Vt = 5·0 V. (peak)
Vr = Volts (peak)

COMPONENT	MEASURED	CALCULATED
D.C.	1·2	1·2
f_1	2·9	2·9
f_2	0·86	0·79
f_3	0·032	0·035
f_4	0·17	0·15
f_5	0·014	0·018

Conclusion

The equivalent circuits described in the foregoing paragraphs are based entirely on empirical values and bear little relation to the theoretical equations relating current and voltage in point contact or barrier layer rectifiers. However, the practical results obtained by their use, within the continuous rated range of a given sample, justify their employment in engineering problems. It is also evident that by a process of trimming and padding an adequate similarity can be obtained among a number of production samples.

The hyperbolic sine function suggested as a relation between the applied voltage and the voltage developed across the 'variable resistor' allows reasonably simple expressions to be derived for rectifiers employed in either symmetrical or non-symmetrical voltage dividers. In spite of the divergent nature of the inverse form of this function a reasonably accurate estimate of the harmonics produced in such a voltage divider can be made by employing the methods indicated above.

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A Transistor D.C. Chopper Amplifier

By P. L. Burton*, B.Sc.

Transistors are used both as switches and as linear a.c. amplifiers to produce a gain-stabilized, low-drift d.c. amplifier for use with thermocouples.

The use of a transistor as a chopper and as a synchronous rectifier is described in detail, and a design procedure given.

Performance figures are quoted for the amplifier, and stress is laid on its economy of space, weight, and power, and on its robust construction, which makes it particularly suitable for use under conditions where mechanical vibration or shock may be encountered.

A LOW drift d.c. amplifier was required to raise the 50mV output of a thermocouple to a level suitable for feeding a multi-channel telemetry transmitter. Only a very limited bandwidth was needed and this seemed an obvious application for a chopper type amplifier, except that the vibration which the equipment would have to endure was too high for mechanical vibrators to be used.

Electronic choppers or modulators were not at the time a very attractive proposition: the diode bridge arrangement—at least before silicon diodes were freely available—was not favoured because it could not be adjusted to give negligible switching 'pedestal' over the considerable temperature range involved, and so a system using a magnetic amplifier was devised.

The advent of the junction transistor and the work done on it by the writer's colleagues to produce a very satisfactory electronic switch radically altered the picture: it has been found possible to produce a chopper device substantially free from pedestal, and furthermore one which is simple, designable, and requires no setting up whatsoever. The complete amplifier occupies only a fraction of the space, and consumes but a fraction of the power, of the magnetic amplifier. Its drift is less than 1mV at the input over a wide range of temperatures, and severe shock, vibration, and climatic testing have no effect on its performance.

The Transistor Switch

The focal point of interest in the amplifier is the transistor chopper, and this will be described in detail first.

The principle of the device can be seen from the conventional characteristic curves of a junction transistor—a typical set is shown in Fig. 1, where collector to emitter voltage is plotted against collector current for several values of base current (i.e. current flowing out of the base electrode).

Each curve has a pronounced 'knee' below which the emitter-collector voltage is a very small fraction of one volt. A transistor operated so that its working point falls below this knee is said to be saturated or bottomed.

The emitter and collector of a junction transistor differ only in size, and it is possible to interchange the functions of the two and still obtain transistor action. The curves of Fig. 2 are obtained in the same way as those in Fig. 1 except that the emitter and collector terminals are interchanged; there is no basic difference in the characteristic, the well-defined knee is still there, and the emitter-collector voltage in the bottomed region is again very small. The only distinctive feature of this reverse connexion is that the base current corresponding to any given emitter current in the bottomed region is much larger than in Fig. 1.

A transistor thus has the following important properties: whether current flows from the emitter terminal to the collector terminal or vice versa, a working point can be chosen where the emitter-collector potential is very small. Alternatively the flow of current can be interrupted if both the emitter and collector diodes are cut off by biasing the base positive with respect to both other electrodes. The emitter and collector terminals can thus form a switch which is controlled by the base electrode: the terminals are open-circuit when the base is biased positively with respect to them, and become a short-circuit (i.e. adopt almost equal potentials) when the base is given a negative bias to produce sufficient outward-flowing base current to bottom the transistor.

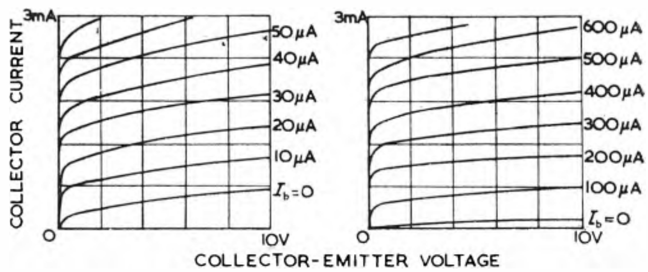


Fig. 1. Normal characteristics of OC71 — Fig. 2. Reverse characteristics

Detailed Design of a Transistor Switch

The chopper circuit based on the transistor switch is shown in Fig. 3 and its extreme simplicity is at once striking. The action is straightforward: so long as the base is held more positive than both the emitter and collector the transistor is off, and the signal is transmitted without loss from A to B, but when the control potential takes the base negative with respect to the collector a base current, defined by R_b, flows, causes the transistor to bottom and point B to fall to earth (collector) potential. Thus if the control potential comprises a train of negative pulses, the signal is chopped in sympathy with the pulses.

Once the resistor R_s has been chosen the peak emitter current that the chopper transistor has to handle is defined by the peak input signal. The maximum collector-emitter potential which can be tolerated when the transistor is bottomed (i.e. the maximum departure from the ideal short-circuit) will in general be fixed by the size of the signal to be chopped and by the precision required. Only one quantity therefore remains to be decided: the base current. The curves of Figs. 1 and 2 are not very convenient for doing that, and so the writer and his colleagues have developed a novel method of presenting the transistor characteristics, shown in Fig. 4.

The collector-emitter potential is plotted against base current with emitter current as parameter. The curves marked I_e denote current into the emitter terminal (i.e. for positive inputs to the chopper), and those marked I_{er}, current out of the terminal (i.e. for negative inputs). With the aid of these curves the correct base current to suit any condition is easily read off.

* *English Electric Co. Ltd.*

In many applications the base current can be defined without any reference to transistor characteristics. This is because the collector-base potential for low power transistors rarely exceeds one volt under bottomed conditions, and thus, provided that the chopping waveform amplitude exceeds two or three volts, all that is necessary is to include a series base resistor R_b such that

$$R_b = \frac{\text{Peak negative excursion of chopping waveform relative to collector}}{\text{Desired base current}}$$

The curves of Fig. 4 show clearly what an excellent switch the transistor is. With zero emitter current, which corresponds to zero signal input to the chopper of Fig. 3, the collector-emitter potential V_{ce} is only about 1mV (point A, Fig. 4), so that, considered at any but very low levels, the device is virtually free from switching pedestal.

An interesting point is that between the lines XX and YY current flows into both emitter and collector simultaneously,

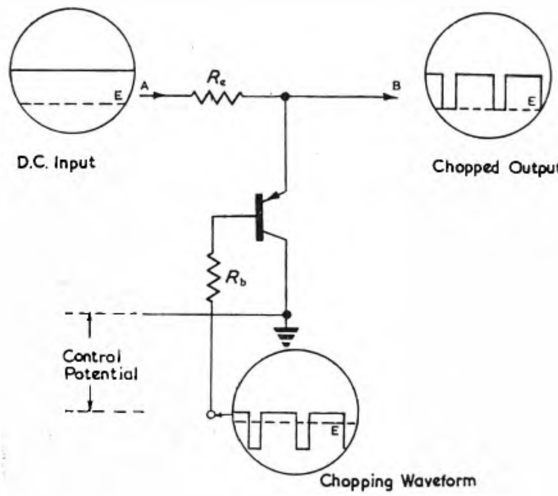


Fig. 3. Basic transistor chopper

a condition completely unlike anything which occurs in the linear operating region. However, the transition from the more familiar state where current flows from emitter to collector or vice versa is perfectly smooth.

The curves of Fig. 4 are not, at present, published by any transistor manufacturer. This may well be remedied in the future, but meanwhile it is a simple, though tedious, matter to determine the curves for a particular transistor type. A series of such measurements including the effect of temperature and scatter among various transistor samples was made some two years ago by the writer and his colleagues, and the work has amply justified itself now that the transistor switch is being widely used by them.

Leakage Current, and a Compromise

The attenuation produced by the chopper circuit when the switch is closed is readily found from Fig. 4. Suppose a signal input of 1V peak is to be chopped, and R_c (Fig. 3) is chosen to be 1kΩ. The peak current through the transistor is then 1mA, and the peak residual output from the chopper when the switch is closed, i.e. the departure from the ideal, is given for a base current of 1mA by the lines AB (for positive inputs) and AC (for negative inputs). Raising R_c to 10kΩ reduces the peak current of 0·1mA, and the residual output to the much smaller values of AD and AE. To reduce the residual output to insignificance it seems simply necessary to go on increasing R_c, but there is a limit. So far only one half of the chopping cycle has been considered in detail; the part of the cycle when the transistor is cut off unfortunately imposes stringent conditions on the choice of R_e.

When the transistor base is held positive with respect to emitter and collector, leakage currents flow outwards in both these electrodes. These currents which the writer denotes, somewhat unconventionally, I_{eo} and I_{co}, are both roughly exponential functions of temperature, doubling approximately for each 10°C rise in temperature. The current I_{eo} flows in the resistor R_e and produces a spurious

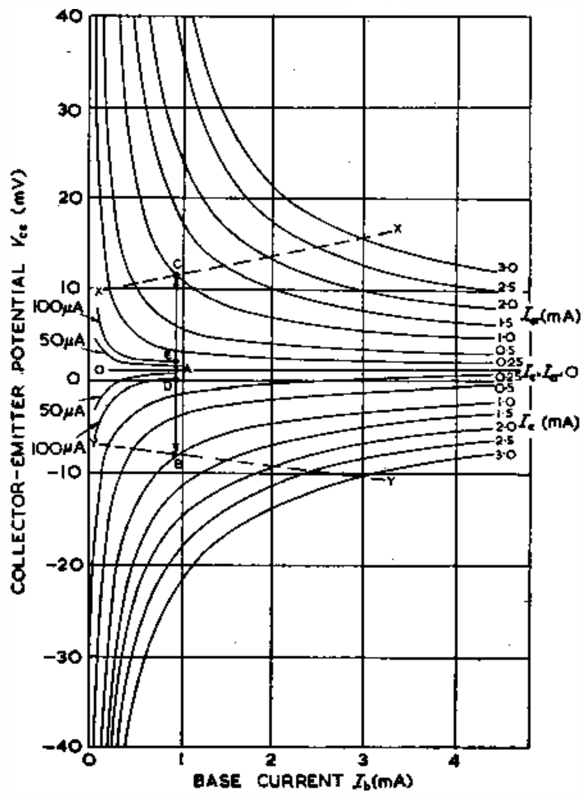


Fig. 4. Showing how residual voltage across switch (V_{ce}) varies with base current and emitter current

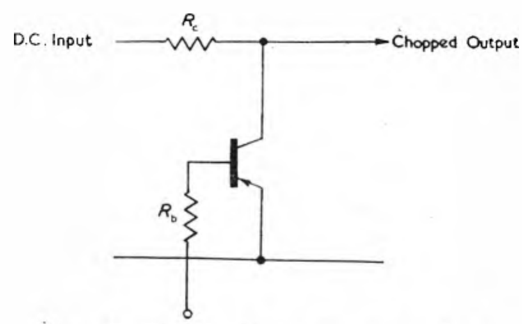


Fig. 5. Transistor chopper—reverse connexion

signal $I_{eo}R_e$ at the output of the chopper. Thus for zero input to the chopper the output fluctuates between $I_{eo}R_e$ (transistor cut off) and V_{ce} (transistor bottomed), i.e. the magnitude of the pedestal is $I_{eo}R_e + V_{ce}$. A compromise is therefore necessary, since for best performance half the cycle demands a large R_e and the other half a small R_e.

In equipment which has to cover a wide ambient temperature range, the design is almost entirely dictated by the $I_{eo}R_e$ term, because I_{eo} is much more temperature-dependent than V_{ce}. For example, I_{eo} in the Mullard OC71 may reach

$10\mu A$ at 50°C compared with $1\mu A$ at 20°C. With R_e equal to 1kΩ, this yields an error of 1 to 10mV, whereas although V_{ce} for 1V input to the chopper is about 10mV (AB or AC in Fig. 4) it does not change a great deal with the temperature. Under laboratory conditions where temperatures are moderate and fairly constant V_{ce} and $I_{eo}R_e$ assume more nearly equal importance.

Symmetrical Transistors

It is a point of great interest that the performance of the switch using a transistor such as the OC71 is seriously degraded if the emitter and collector are reversed as in

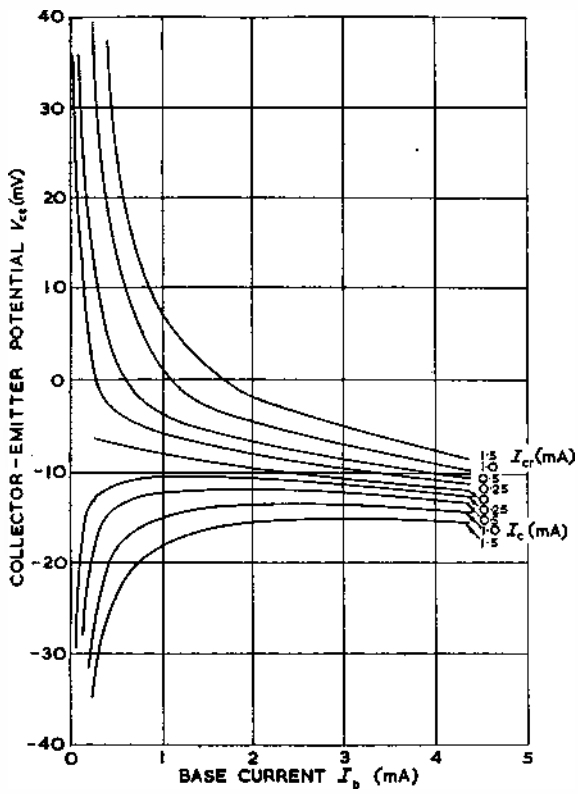


Fig. 6. The data of Fig. 4 replotted with collector current I_c or I_{cr} as parameter

Fig. 5. There are two reasons for this: the spurious output due to leakage current when the switch is open becomes $I_{co}R_e$, undesirable since typically I_{co} is five times larger than I_{eo}; and V_{ce} for zero input current becomes much larger. This latter point is brought out in Fig. 6 where the same information given by Fig. 4 is replotted with collector current as parameter, using the relation $I_c = I_e - I_b$. The $I_c = 0$ curve is seen to lie well away from the $V_{ce} = 0$ axis.

A useful conclusion can be drawn from this result: any transistor designed specifically for this type of switching should not, as is often advocated, be symmetrical. For a perfectly symmetrical transistor (i.e. one with identical emitter and collector junctions) Figs. 4 and 6 would be identical; consequently in neither plot could the $I = 0$ curve lie close to the $V_{ce} = 0$ axis.

The Synchronous Rectifier

The transistor switch is used again in the output circuits of the amplifier as a synchronous rectifier, driven by the same waveform as the chopper. The circuit is shown in Fig. 7 together with the relevant waveforms.

At the relatively high level which exists here, the waveform is rather over 3V peak-to-peak, there is no need to study the variation of V_{ce}. The leakage current I_{eo}, however, is still important since when the switch is open the capacitor C charges up. Now I_{eo} is roughly constant and independent of emitter-base voltage, so that the output voltage rises at I_{eo}/C volts/sec during one half of the cycle, and falls almost to zero during the other. This sawtooth output wave appears after smoothing by the circuit C_sR_s as a steady zero error dependent upon ambient temperature. It can be shown that if the clamp is fed from an amplifier of output impedance R_o, and CR_o is large compared with one cycle of the chopping wave, then the output error is

$$nI_{eo}R_o$$

where n is the mark-to-space ratio of the chopping wave.

Choice of Chopping Waveform and Frequency

The mark-to-space ratio of the chopping waveform in the amplifier is the result of a compromise. For unity ratio

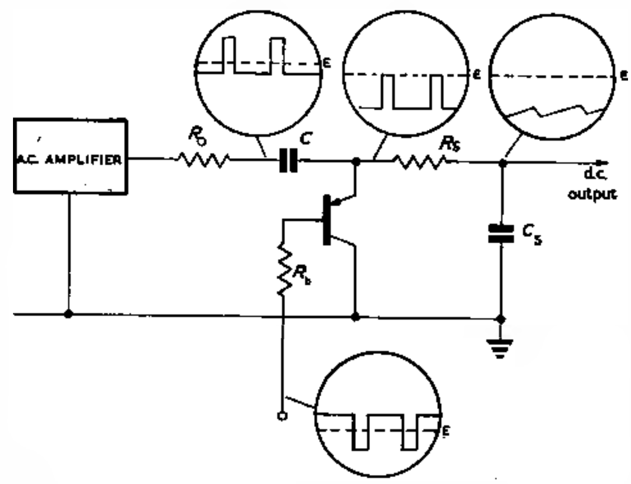


Fig. 7. Synchronous rectifier or clamp

the a.c. amplifier and the clamp must handle a peak-to-peak signal amplitude of twice the smoothed d.c. output; for a ratio of 10 or more the two are almost equal, but the clamp zero error is greater. Since the signal handling capacity of transistor amplifiers is limited, and the output is fixed at 3V d.c., a figure of approximately 10 is chosen. A further increase would only slightly reduce the peak a.c. signal, an improvement which would not be worth the increased clamp zero error.

The choice of chopping frequency depends on what frequency response is required from the amplifier. In the present case, only a very limited bandwidth was required, and the choice of chopping frequency was not very critical.

A multi-stage a.c. amplifier, using an audio-frequency transistor, such as the OC71, cannot conveniently handle pulses of less than 100μsec duration, which, with a 10/1 mark-to-space ratio, sets the upper frequency limit at about 1kc/s, and this was the value chosen here.

The A.C. Amplifier

A substantial amount of feedback is applied over-all to the chopper amplifier, and consequently an a.c. amplifier with a large gain is required. The amplifier is within the feedback loop and its stability at both high and low frequencies must be considered.

High frequency stabilization of a transistor feedback amplifier is usually a very difficult problem, because in addition to phase-shifts produced by stray capacitances, there are phase-shifts produced by the transistors themselves. However, in this case, there is little difficulty since it is possible to make the lagging time-constant formed by C, C_s, and R_s (see Fig. 7) the major time-constant in the loop. Its value can easily be chosen to give an attenuation greater than the loop gain at frequencies where the loop might otherwise become unstable.

The low frequency stability problem is overcome by including only two time-constants in the loop.

The Thermocouple Amplifier in Detail

Fig. 8 shows the complete circuit of the thermocouple amplifier. The thermocouple is connected directly to the amplifier input, so that its low internal resistance (about 10Ω) forms the input resistor, in this way the zero error arising from leakage current is kept as small as possible.

with the smoothing circuit $R_{15}C_7$. The clamp transistor X_7, like the chopper X_1, is designed to have a base current of 1mA in the conducting state, and is driven from the same switching waveform. The filter for smoothing the output, $R_{15}C_7$, feeds into the first of two cascaded emitter-followers X_8 and X_9, which transform the output impedances of the rectifier filter system (several thousand ohms) to a value of a few tens of ohms, suitable for driving the feedback resistor R_{25}.

The open loop gain of the chopper amplifier system is about 1 000, reduced to about 55 with feedback applied. As well as its normal function of stabilizing the gain of the amplifier, the feedback reduces the current flowing through the chopper transistor X_1, and with it the part of the pedestal due to V_{ce}.

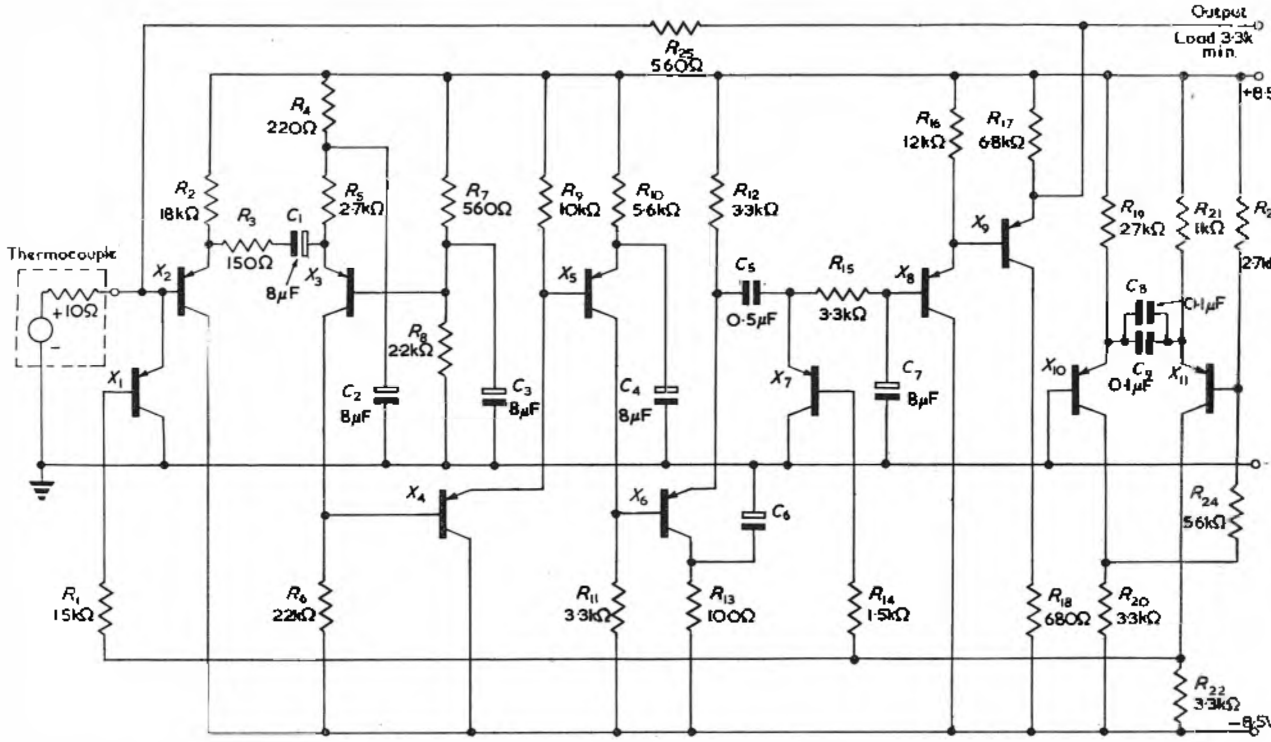


Fig. 8. Circuit diagram of thermocouple amplifier

The multivibrator $X_{10}X_{11}$ provides a switching waveform at about 1·2kc/s for the chopper X_1, which is made to conduct for approximately one-eighth of each cycle. The base current of X_1 is defined at 1mA during the conducting period by resistor R_1. The chopper is coupled directly to the a.c. amplifier, which comprises X_2 to X_6. There are only two stages of amplification: the emitter-coupled pair X_2, X_3, and the grounded emitter amplifier X_5. X_4 and X_6 are emitter-followers, which by reducing the shunting effect of the succeeding stages on collector loads R_6 and R_{11}, allow much larger gains per stage than would otherwise be possible. Direct couplings are employed from X_3 to X_6 inclusive, to avoid the phase-shift of capacitor couplings. To ensure stability of the working point of the transistors in this part of the amplifier, currents in X_3 and X_5 are closely defined by the emitter resistors (R_4+R_5) and R_{10}.

The only two circuits producing phase lead are formed respectively by C_1 with R_3 and the input impedances of the emitters of X_1 and X_2, and by C_4 with the input impedances at the emitter of X_5. Capacitor C_3 gives no additional phase lead, since electrically its effect is equivalent to a slight reduction in the capacitance of C_1.

The synchronous clamp is formed by C_5 and X_7, together

Performance

The production model of the amplifier makes use of printed wiring techniques, and all the components, including the eleven transistors, are encapsulated in Araldite resin, the whole structure occupying the space of a 2in cube.

The characteristics of this amplifier are almost constant over a range of ambient temperature from −12°C to +50°C. The total change in the zero error in this range is less than 1mV at the input.

The amplifier draws only 330mW from its supplies of plus and minus 8·5V, and is very little affected by supply changes. A 30 per cent reduction in the voltage causes a change of gain of less than 1 per cent, but reduction of both supplies below 7·5 volts limits the maximum output to less than 3V.

Shock, climatic, and vibration tests have been applied to the amplifier, but none has had any effect on its performance.

Conclusion

The use of transistors as switches enables a high gain, all-transistor chopper amplifier to be designed, which will handle inputs of tens of millivolts, and which is virtually drift-free over a large range of temperatures.

At the low power level at which such an amplifier operates, very little heat is dissipated, so that the components may be mounted close together, and the instrument made physically very small.

Printed wiring and 'potting' techniques are ideally suited to the construction of this amplifier, and result in the production of a robust, compact unit, extremely useful in applications where saving in space, weight, and power consumption are of importance, but where a precision instrument is nevertheless required.

Acknowledgments

The writer thanks Mr. L. H. Bedford, Chief Engineer of the English Electric Co. Ltd., Guided Weapons Division, for permission to publish this article, Mr. A. J. Vaughan, now Assistant Chief Development Engineer, for much help with design and development of the amplifier, and Mr. T. K. Hemingway for his painstaking experimental work.

A Doppler Navigator

It is now possible to disclose some details of the hitherto secret Marconi Doppler Navigator Type AD.2000, which has been in quantity production for the Royal Air Force for the past three years.

Briefly, the AD2000 is a completely self-contained airborne equipment which provides an automatic and continuous flow of navigational information such as immediate position in latitude and longitude, track guidance, distance run or distance to go, estimated time of arrival and wind velocity.

The overwhelming advantages of such an airborne navigational aid for military operations are obvious. It is the only system in full production today which does not depend upon the co-operation of ground-based stations and therefore the only one (apart from the use of astro-navigation) which is capable of independent global operation.

The Navigator type AD2000 measures the ground speed and drift angle of an aircraft in flight by making use of the 'Doppler Effect.' Electromagnetic waves transmitted from the aircraft strike the ground and are deflected in various directions, a small portion of the energy returning to the aircraft. Due to the relative motion of the aircraft to the ground the frequency of the return signal differs slightly from the transmitted one. This difference in frequency bears a direct relationship to the ground speed of the aircraft.

In the AD2000 the antenna system radiates two beams, one forward and the other backward, both beams being depressed at an angle toward the ground. The ground speed is found by measuring the beat frequency produced when the return signals from the forward beam are mixed with those from the backward-looking beam. (This procedure avoids the necessity for the ultra-stabilization of the transmitter which would be imperative if it was used as a comparison source).

Additionally one beam is displaced to starboard and the other to port, the positions being switched twice a second. The drift angle is found by comparing the doppler frequency obtained when the forward beam is displaced to starboard and the backward beam to port with the frequency when the beam positions are reversed. The antenna is then automatically rotated until the two frequencies are equal. It is then aligned along the aircraft track.

A pulsed system is used, at a p.r.f. of 50kc/s, with a pulse width of 0·45μsec. The pulses are generated by a magnetron which operates in bursts of 40msec duration, twice a second. The peak power is 8kW (mean power, approximately 12W). The operating frequency is within the band 8 750 to 8 850Mc/s.

The equipment consists of the antenna system, transmitter/receiver, tracking unit and the indicator.

The antenna which is unpressurized, consists of four slotted linear arrays lying parallel to each other in a directional horn assembly, the axes of the arrays being horizontal. An arrangement of phased and anti-phased pairs, with a common feed at one end, provides the forward and backward-looking beams. The beam width at the half power points is $2\frac{1}{2}°$.

The two pairs of antennae are energized alternately by a special Y-section of waveguide which oscillates about a vertical hinged joint to form a waveguide switch. The horn assembly deflects one pair of beams to a mean angle of 20° to port and the other pair 20° to starboard.

The waveguide switch is driven at a rate of 1c/s from a small gearbox. Cam-operated switches on the driving shaft provide sense and timing pulses to the remainder of the equipment.

To determine drift the antennae can be rotated 20° either side of the fore-and-aft line of the aircraft. The azimuth gearbox carries a synchro transmitter which repeats the aerial position to the indicating unit.

The transmitter/receiver consists of two pressurized containers. Housed in one of these are the magnetron, the modulator circuits, the receiver klystron and the first i.f. amplifier and a.f.c. circuits.

The second container houses the main h.t. and e.h.t. circuits, an associated delay circuit and relays. A pressurized duct connects the two units beneath the base castings so that e.h.t. (8kV) leads are unbroken and remain within the pressurized containers.

The microwave circuits consist of a duplexer labyrinth connecting the magnetron and klystron to the antenna via a double-directional coupler which provides monitoring facilities for the forward and backward-looking beams.

The modulator is controlled by a crystal oscillator which maintains the p.r.f. at 50kc/s. As stated, the pulses are fed to the magnetron in bursts of 40msec duration, repeated at half-second intervals in synchronism with the movement of the waveguide switch in the antenna system.

The i.f. amplifier, tuned to 45Mc/s is divided between the transmitter/receiver and the tracking unit. The first i.f. amplifier (contained in the transmitter/receiver) is rendered inoperative during the transmitted pulse by the application of a short blanking pulse from the modulator.

The tracking unit is self-contained with its own power supplies within a pressurized container. It contains the frequency-measuring circuits for the determination of ground speed and drift angle, together with the second i.f. amplifier and a.g.c. circuits.

The frequency-measuring circuits consist of two discriminator channels, an integrator circuit, a phonic wheel velodyne and the azimuth drive circuits. In the discriminator channels the doppler frequency is compared with the outputs of two phonic wheels and the resultant frequency is used to control integrator and azimuth drive circuits.

The integrator controls the speed of the motor driving the phonic wheels; this speed is a measurement of the aircraft ground speed.

The azimuth drive circuit controls the movement of the azimuth drive motor situated in the antenna system and rotates the antenna for drift angle measurement. The resultant ground speed and drift information is fed by synchros to an indicating unit.

The indicator unit displays the following information:—

Ground Speed: 100-700 knots, at 5 knot intervals (scale and pointer).

Drift Angle: 20° port-20° starboard (scale and pointer).

Distance Flown: Indicated on a counter to tenths of a nautical mile.

The indicator unit also contains a series of neon lamps and inching controls. The neons give a supplementary approximate indication of ground speed and are used in conjunction with the inching controls to lock the equipment for correct operation. They are also used to indicate when the equipment is locked on signal.

The ground position indicator provides automatic and continuous information of the aircraft's position either in latitude and longitude, as a grid (nautical miles gone N or S and E or W) or as an A/A set track (along and across a set-in track). Choice of presentation is afforded by a switch.

The instrument is primarily a mechanical type of computer employing a conventional disk-ball-roller resolving gear set by the track drive to split incoming ground mileage into two components at right-angles. The component outputs from the resolving gear drive the cyclometer-type counters at the front of the instrument; these show the ground position of the aircraft.

In order that information shall not be lost while setting the counters in flight, a fixing and storing device is incorporated. When obtaining a 'fix' the two output drives from the resolving mechanism are switched from the counters to two sets of storage drums. After adjustment to the counters the switch is turned back to its normal position. The stored mileage is then automatically driven back into the counters, together with the normal incoming mileage.

An H.T. Overload Cut-out Circuit

By J. D. Ralphs, B.Sc.

For the protection of electronic equipment against damage due to overload, a relay cut-out has considerable advantages over the more normal fuse-links in cases where the extra cost is justified. The circuit described was designed in conjunction with a general-purpose stabilized h.t. power supply used for experimental work in a laboratory. Facilities provided include an audible warning of operation of the cut-out, automatic restoration of the supply after a few seconds delay, and 'lock-out' of the supply if the fault still exists at that time.

THE normal method of protecting low power electronic equipment from excessive damage in case of breakdown or overload is to insert one or more fuse-links or thermal cut-outs in series with the primary or secondary windings of the power supply transformer. While this system gives adequate protection for most purposes, there are circumstances in which the higher cost of a more complex overload protection device is justified.

A particular case in which this was considered to be so was in the design of a stabilized d.c. power supply, intended for general purpose laboratory use. In rejecting fuses as the principal form of protection the following points were considered:

(1) The power supply would be in use almost continuously to supply h.t. to experimental equipment, mostly of a temporary nature. Under such circumstances it would be far more subject to overloading and abuse than would be the case with most test gear.

(2) Many cases of overload would consist of accidental short-circuiting of the output h.t. line. Such faults would be easily and quickly 'cleared', but the subsequent replacement of a fuse would be unnecessarily expensive, time-consuming, and irritating to other persons using the same power supply.

(3) In cases where there are a number of heater windings on a power supply transformer, a given percentage increase in h.t. current will cause a much smaller percentage increase in primary current, and in order to give the closest control over the maximum permissible h.t. current the fuses must be inserted in the secondary circuit, where they are subject to considerable peak currents and switching surges with consequent embrittlement and premature failure.

(4) The current at which a fuse will 'blow' is subject to a relatively wide tolerance and varies considerably with the age of the fuse and the time for which the overload persists. For example, a standard 1¼in by ¼in diameter glass cartridge fuse rated at 250mA is specified as blowing 'within 10sec' on a steady current of 500mA. Such a fuse may carry a current of 1A for 0·1sec or more, while failing to blow on a continuous current of 400mA. In an efficient stabilizing circuit the series valve may be run at its maximum rated anode dissipation, and a 70 per cent overload, particularly if inadvertently allowed to continue for some time, could seriously reduce the life of the valve.

(5) The relatively high cost of the electronic stabilizing circuit and the reduction of the possibility of damage to expensive meters, etc., in use on experimental circuits would justify the provision of a more satisfactory protection system.

Facilities

It was evident that several forms of relay cut-out circuit could be designed to comply with the requirements and the one finally developed is described below. It performs the following functions:

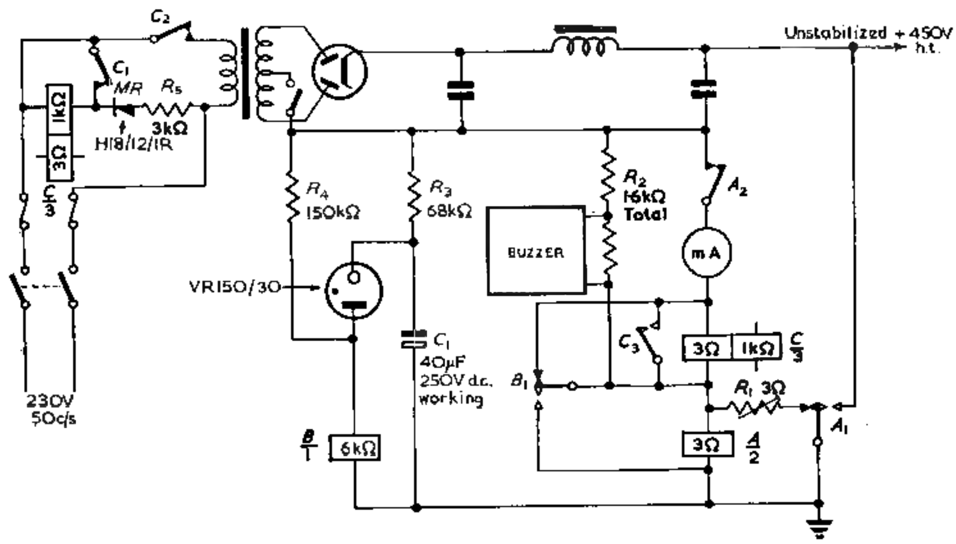


Fig. 1. Overload cut-out circuit

When the total h.t. current rises above a predetermined value the unstabilized h.t. line is short-circuited to earth and an audible warning is sounded. After a delay of about 5sec the supply is restored and, if the fault has been 'cleared', the circuit is back to normal. If, however, the fault persists, the primary circuit of the mains transformer is broken and the circuit 'locks out'. To restore the supply the mains switch is operated off and on again.

The circuit requires no auxiliary supply and may be adapted to be included in most normal h.t. supplies, stabilized or unstabilized. The particular unit for which it was designed delivers 300V heavily stabilized at a nominal maximum current of 300mA.

A disadvantage of the circuit is that it gives no protection against faults in the transformer, rectifier or smoothing circuit, and these are covered by normal fuses in the primary.

Mode of Operation

The circuit (of the protection device only) is given in Fig. 1 and its action may be described in three stages.

CUT-OUT

In the unoperated condition the relay A is shunted by R_1, a preset resistance of about 3Ω maximum, which is adjusted so that if the total h.t. current rises above the maximum allowable, the relay operates. When this occurs, the contact A_2 opens, limiting the fault current; contact A_1 then short-circuits the h.t. line to ground, allowing the full voltage of the supply to be generated across R_2, the dummy load, with, of course, negative polarity. This also operates the buzzer.

DELAY

The negative voltage across R_2 charges the capacitor C_1 through R_3 until the neon striking voltage is reached. When the neon strikes, the relay B is operated and contact B_1 short-circuits relay A, causing it to drop out. When contact A_2 closes, the h.t. current is not passed through relay A (since B_1 is operated) but through the low-impedance coil of relay C, which also has an operating current of approximately the maximum allowable h.t. current. If the fault has been removed during the delay time, the h.t. current is insufficient to operate relay C, and, since contact A_2 has short-circuited the voltage across R_2, capacitor C_1 discharges through R_3, relay B drops out and the circuit is restored to normal. Relay B will actually release almost immediately after the removal of its supply by contact A_2, and any subsequent overload will be cleared in the way described above, except that, if C_1 has not completely discharged, the time delay will be shorter.

LOCK-OUT

If at the end of the delay time the fault has not been cleared, when contact A_2 closes the current will be sufficient to operate relay C. Contact C_2 then breaks the mains supply to the transformer primary, while contact C_1 'locks-in' relay C by means of its own half-wave supply across the mains. The h.t. supply may be restored by breaking the mains supply momentarily by means of the 'mains on-off' switch. This causes relay C to drop out and conditions are returned to normal.

Design Considerations

The more detailed design may now be considered as follows:

CUT-OUT

It is evident that to ensure a positive action of relay A on slowly-rising fault currents, contact A_1 must open (increasing the sensitivity of the relay) before contact A_2 opens. However, during the transition of contact A_1, the total load on the power supply consists of R_2 in series with the external load (see Fig. 2), and the further condition must be observed that the current under these conditions must still be greater than the unshunted operating current of the relay. To avoid overloading the power supply during the delay period, the minimum value of R_2 is that which, with a mains voltage equal to the maximum design limit, will pass slightly less than the maximum permissible current. The maximum value of R_2 is that which, in series with an external load just great enough to trip the cut-out, and with minimum mains voltage, will pass sufficient current to complete the pull-in operation of relay A. It would seem then, that the best course would be for relay A to be relatively sensitive, i.e. to operate on, say, 10 per cent maximum permissible current. However, this has two disadvantages:

(1) The impedance of the relay will be relatively high, thus increasing the output impedance of the unstabilized section of the power supply. In this particular case, this is immaterial, but in a power supply delivering a lower current the relay impedance may be of several hundred ohms.

(2) A relatively larger proportion of the current will be passed through R_1 and the contact A_1.

On the other hand, a low impedance relay will require a low impedance shunt, in which case any variations in contact resistance of A_1 may cause variations in the operate current of the relay. The compromise arrived at in this instance was that relay A should operate on slightly less than half the required cut-out current (i.e. about 120 to 130mA), and the current through R_2 in the fully operated condition is about 260mA.

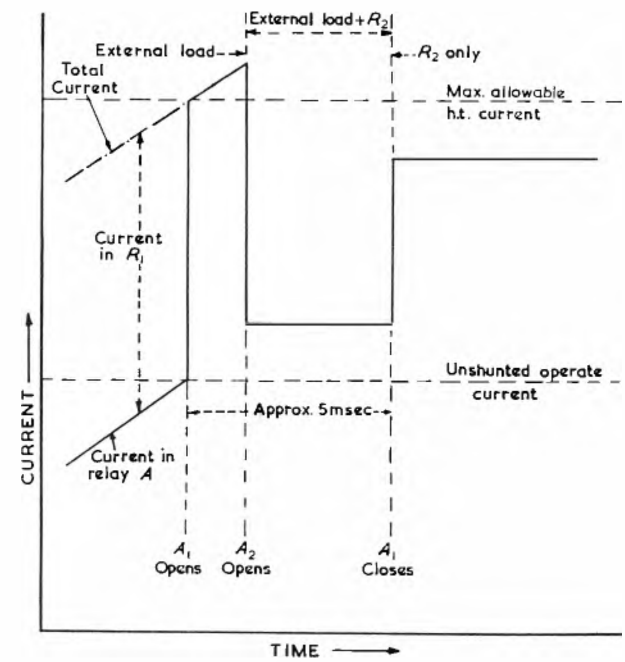


Fig. 2. Operation of relay A

The negative terminals of the smoothing capacitors must be insulated from the chassis to withstand the full h.t. voltage. In the interests of reliability of the equipment as a whole electrolytic capacitors were avoided where possible and these capacitors are of the paper type, and therefore this requirement is easily met.

There is an alternative circuit which may be considered if electrolytic capacitors are used, as it does not require the 'cans' to be insulated. In this arrangement the negative terminals of the capacitors are earthed and the 'make' contact of A_1 is returned to the rectifier cathode. This circuit is not recommended, however, for two reasons. Firstly, it requires that the contact A_1 shall, on the operation of the relay A, discharge the reservoir capacitor, and at the end of the delay time, when relay A releases, the full charging current of the smoothing network shall pass through A_2. Since both contacts normally make or break a large proportion of the maximum h.t. current, the addition of these extremely heavy surge currents considerably increases the burden on them. With the circuit shown, the contacts are operating into virtually resistive loads and there is no large change in the voltage across the smoothing circuit.

The second advantage of the circuit of Fig. 1 is that, as soon as A_2 opens, the current through the external fault is restricted to about half the maximum allowable, and on the completion of the action of relay A (normally about 5msec) the output is short-circuited. With the alternative circuit the voltage across the reservoir capacitor is removed, but the other smoothing capacitors will continue to discharge through the external fault, and the possibility of damage to meters, etc., is considerably increased.

Delay

It is obviously advantageous to keep the value of capacitor C_1 to the lowest possible value, particularly if this enables a paper capacitor to be used (see below). For a given relay operating current, the major parameters controlling the delay (apart from the capacitor itself) are the resistor R_3 and the striking voltage of the neon (the effect of R_4 is considered later). Due to the inductive impedance of the relay, the discharge of the capacitor cannot contribute appreciably to the current operating it, and R_3 must be low enough to pass the required current under steady-state conditions with the mains voltage at the lower design limit.

A rough estimate of the optimum striking voltage of the neon may be obtained as follows:

Assume that the neon striking voltage and maintaining voltage are the same, and that the resistance of the relay is small compared with the series resistor R_3. (Both reasonable first-order approximations).

From Fig. 3:

If E = Voltage generated across R_2 with minimum mains voltage

V = Maintaining voltage of neon

I = Required relay current

T = Time delay

$R = R_3$

Then $(V/E) = 1 - e^{-T/CR}$

and $E = V + IR$

$\therefore T = -CR \ln (RI/E)$

$dT/dR = -C(1 + \ln (RI/E))$

$dT/dV = (dT/dR) \, . \, (dR/dV) = (C/I)\,(1 + \ln(RI/E))$

The condition for maximum delay is therefore

$\ln (RI/E) = -1$ i.e. $V/E = (1 - (1/e)) = 0{\cdot}63$

and the delay time is then given by

$T = CR.$

However, the maximum is broad and values of V/E between 0·4 and 0·85 give delays within 25 per cent of the theoretical maximum. In practice, since the striking voltage of the neon is higher than the maintaining voltage, the delay will be somewhat greater than the theoretical value.

In the unit under consideration E was 420V and unless two neons in series were used the value of V/E was restricted to below about 0·35—given by a neon type OD3/VR150.

The effective operating current of the relay may be reduced by the current flowing in R_4 and the value of R_3 correspondingly increased. The current in R_4 may be greater than the drop-out current of the relay, since, when relay B operates and releases relay A, the energizing voltage B is short-circuited by A_2. The upper limit is set only by the stability of the relay B operating current and the necessity for the by-pass current to remain below this value when the mains voltage is at its design maximum.

It was indicated above that it is preferable for capacitor C_1 to be a paper type. This was not found practicable for any relay available. One effect of using an electrolytic capacitor is that if the cut-out is not operated for some time, the first delay is much longer than the 5 to 10sec normally obtained. This increase is presumably due to the 'forming' current of the capacitor reducing the voltage across the neon.

Lock-out

The operating current of relay C by the low impedance coil is chosen (or adjusted by shunting) to be slightly less than that of relay A to prevent any possibility of 'cycling' of relays A and B under marginal fault conditions.

The 'lock-out' circuit has been designed to operate on a 'break' contact C_1, rather than a 'make', since with the particular relay used it is easier to adjust contacts C_1 and C_2 to energize the high impedance winding before the mains supply is broken. Since this winding is connected so as to aid the current in the low impedance winding, a fast and positive action is ensured. It has been mentioned that, in the interests of reliability, electrolytic capacitors have been avoided where possible in the equipment as a whole, and it

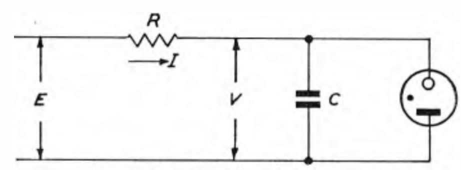


Fig. 3. Delay circuit

was found that no smoothing is necessary in this circuit if the contact C_3 is used to short-circuit the low impedance winding and effectively 'slug' the relay. This arrangement also allows a smaller rectifier to be used, since it reduces the peak inverse voltage generated across it.

Components

The unit itself was constructed from the limited components available, and the values given should be considered merely as indicating the orders to be expected. This applies particularly to the relays, which are all of the Post Office type 3000, but which were rewound where necessary to give the required operating currents. Relays B and C (high impedance) could operate on lower currents to some advantage.

Care should be taken that the high impedance coil of relay C and the contacts C_1 and C_2 are sufficiently well insulated from the other coil and from the yoke. On relays A and C high voltage heavy duty Siemens 'H' type spring sets were used to give a high reliability and safety factor, although, as indicated above, those contacts making or breaking heavy currents do not operate into highly reactive loads.

In view of the relatively high speed of operation mentioned, it was considered that a special high-speed relay was unnecessary even if one were available with the necessary contact capacity.

The 'audible warning' consists of a 1kc/s 'microphone hummer' (as used in some battery-operated a.c. bridges), capacitively coupled to a telephone receiver earpiece mounted on the front panel, the low voltage d.c. required being obtained by a tap on the dummy load chain. The resulting note, while hardly euphonious, is adequate.

The dummy load itself is required to dissipate about 120W, but since this is only carried for a few seconds, two wire-wound resistors of 60W total nominal rating are used. Provided that they are well ventilated and mounted clear of other components, this would seem to be a suitable arrangement. R_5, a wire-wound resistor nominally rated at 10W, should also be well ventilated.

LETTERS TO THE EDITOR

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

A Square Wave Convertor

DEAR SIR,—Having recently developed a circuit similar to that described by Mr. J. B. Earnshaw in your April issue, I should like to draw attention to the conclusion drawn from equation (5), i.e. that the operation of the circuit is dependent on backlash in the regenerative squarer.

Equation (5) states that

$$\frac{dp}{de} = (1/T).(\cot\ \psi_1 - \cot\ \psi_2)$$

the condition, $\cot\ \psi_1 - \cot\ \psi_2 = 0$ can only be satisfied when $\psi_1 = \psi_2$ and the only occasion on which this can occur in a practical waveform is when $\psi_1 = \psi_2 = 90°$. The condition of no backlash, i.e. when $\psi_1 + \psi_2 = 180°$ implies that $\cot\ \psi_1\ 0\ \cot\ \psi_2 = 2\cot\ \psi_1$ which shows that the effectiveness of the bias control on mark-to-space ratio is not lost.

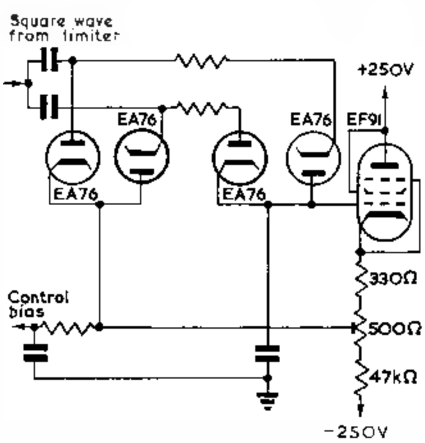


Fig. A. The control circuit

The circuit developed by the writer is a precision limiter with its mark-to-space ratio held at unity by a feed-back arrangement. The limiter contains no regeneration and operates with negligible backlash, while bias control of the mark-to-space ratio is entirely effective. The control part of the circuit is shown in Fig. A, the error detector is very similar to Mr. Earnshaw's, while the d.c. amplifier, which was tried and found to suffer from drift and susceptibility to supply voltage changes, was replaced by a bootstrap integrator. Following the integrator some additional smoothing is required, the complete system forming a 'zero displacement error' servo, the integrator and smoothing time-constants must be chosen to give stability. The bootstrap integrator, being essentially a cathode-follower, is stable and relatively unaffected by the supply voltages.

The writer is indebted to the Admiralty for permission to publish these details.

Yours faithfully,

H. E. HYDE,

Admiralty Research Laboratory.

The author replies:

DEAR SIR,—I would like to thank Mr. H. E. Hyde for pointing out an error in the solution of equation (5) in my article 'A Square Wave Convertor'. The correct solution of $\cot\ \psi_1 - \cot\ \psi_2 = 0$ is of course $\psi_1 = \psi_2 \pm n\pi$ and the incorrect solution $\psi_1 + \psi_2 = 180°$ should be deleted.

Fortunately this error has little effect on the rest of the argument for although there appears to be control when the backlash is zero, i.e. $dp/de = 0$, this control cannot be utilized since the fast squaring action of the Schmitt trigger ceases and the circuit no longer operates in the intended manner.

The circuit of the convertor as published, showing an uncompensated d.c. amplifier in the feedback loop, was used experimentally for short periods only, and I agree that for long term usage the amplifier must be compensated for supply voltage variations.

However, I most certainly disagree with the common fallacy that cathode-follower type circuits are relatively unaffected by variations in the supply voltage. Negative feedback, while stabilizing the gain of an amplifier to which it is applied, does not in any way compensate for variation in the output voltage which originate due to variations of the supply voltage.

For, consider a valve operating with anode load R_L and cathode load R_K, this represents (1) a simple amplifier when $R_K = 0$, $R_L > 0$; (2) an amplifier with negative current feedback when $R_K > 0$, $R_L > 0$; and (3) an amplifier with negative voltage feedback when $R_K > 0$, $R_L = 0$; and the output is taken from the cathode, i.e. a cathode-follower.

If a change in the supply voltage produces a change of ΔE_a in the h.t. voltage and a change of ΔE_h in the heater voltage, the corresponding increase in anode current ΔI_1, is:—

$$\Delta I_1 = \frac{\Delta E_a}{R_L + r_a + (\mu+1).R_K} + \frac{(\mu+1).k.\Delta E_h}{R_L + r_a + (\mu+1).R_K} \quad \ldots\ldots\ldots\ldots (1)$$

where k is the ratio of the heater-anode, transconductance to the grid-anode transconductance[1].

The increase in anode current ΔI_2 brought about by an input signal voltage of ΔE_g is:—

$$\Delta I_2 = \frac{\mu.\Delta E_g}{R_L + r_a + (\mu+1).R_K} \quad \ldots\ldots\ldots\ldots (2)$$

and for ΔI_1, and ΔI_2 to be equivalent

$$E_g = 1/\mu.\Delta E_a + \mu + 1/\mu.k.\Delta E_h \quad \ldots (3)$$

so that ΔE_g is then the input signal voltage to which the supply voltage variations are equivalent, and from which they cannot be distinguished.

Briefly, the current changes produced by the supply voltage variations are reduced when negative feedback is applied to an amplifier (see equation (1), $R_K > 0$), but so are the current changes produced by the input signal voltages, and by exactly the same factor.

It would appear then that the advantages claimed by Mr. Hyde for his circuit must be due to the different method of obtaining and applying the error signal, since the bootstrap amplifier is as susceptible as any other uncompensated amplifier to supply voltage variations.

Yours faithfully,

J. B. EARNSHAW,

Auckland University College.

REFERENCE

1. EARNSHAW, J. B. Heater Voltage Compensation for DC Amplifiers. *Electronic Engng, 29,* 31, (1957).

PUBLICATIONS RECEIVED

OPPORTUNITIES IN ELECTRONICS is the title of a booklet issued by Mullard Ltd, Mullard House, Torrington Place, London, W.C.1. It shows the science graduate or student what the Mullard electronics organization has to offer to the young person seeking a career in electronics.

TRANSISTORS CIRCUITS AND SERVICING by B. R. A. Bettridge explains in simple practical terms how transistors work, how they are used in radio circuits, and the best methods to employ when servicing equipment that uses them. The treatment is almost entirely descriptive, mathematics being avoided, and the technical level is exactly right for the service engineer. Trader Publishing Co. Ltd, Dorset House, Stamford Street, London, S.E.1. Price 2s. 6d.

THE EFFECTS OF RADIATION ON ORGANO-POLYSILOXANES is the title of the most recent publication issued by Midland Silicones Ltd in their series of Silicone Notes. The publication is divided into two major sections, one of which deals with the use of irradiation techniques for curing polymers, while the other describes the ageing effects of radiation on conventionally cured silicone products. Copies (reference number Silicone Notes A 8) are obtainable on request from Midland Silicones Ltd, 19 Upper Brook Street, London, W.1.

ENGINEERING TRAINING IN THE BBC by K. R. Sturley is No. 11 in the BBC Engineering Monograph series. Individual copies, post free, are available from BBC Publications, 35 Marylebone High Street, London, W.1, Price 5s.

REPORT OF THE MEETING ON SEMICONDUCTORS held by the Physical Society in collaboration with the British Thomson-Houston Co. Ltd of Rugby in April 1956, has been issued by the Physical Society, 1 Lowther Gardens, Prince Consort Road, London, S.W.7. Price 12s. 6d. to members and 20s. to non-members.

BOOK REVIEWS

Analog Computer Techniques

By Clarence L. Johnson. 264 pp. 50 figs. Demy 8vo. McGraw-Hill Book Co. 1956. Price 45s.

THIS book sets out, in the words of its author, "to shorten the period of transition from neophyte to accomplished analog-computer operator for the engineer learning to use this new tool." It is also stated that the "presentation is such that the average person with a knowledge of Ohm's Law, Kirchoff's Laws, and a basic knowledge of differential equations can read and understand the majority of the material presented."

The reviewer's reaction on reading the early chapters was that the pace was much too rapid for a beginner defined as above. In particular, basic concepts, such as the use of a closed-loop arrangement to solve a differential equation, were too briefly dealt with. As a book for beginners more care should also have been exercised in the definition of terms and the clarification of ideas with which the beginner may be assumed to be unfamiliar. All too often the definition or clarification appears several chapters too late. For example, the "stabilization" of amplifiers, the "sign-reversal" associated with computing amplifiers and the "transfer function" are all frequently used in the early chapters without comment and with explanation belatedly appearing in later chapters. "Repetitive computers" are contrasted with "real time computers" whereas they should be contrasted with "single shot computers" irrespective of time scale.

No detailed electronic circuits are discussed. It is simply assumed that electronic units exist which can, by analogue means, carry out the basic mathematical operations of addition, integration, multiplication, etc., and attention is concentrated on the techniques whereby such units can be inter-connected and combined with diodes or relays to yield solutions to equations, primarily non-linear integro-differential equations, arising in mathematical and physical problems. This is a sound introductory approach but it has its dangers. Almost without exception it is the application of complete and commercially available American machines, such as the REAC, which are discussed and the use of these as examples leads in certain cases to unqualified statements which are wrong or misleading when applied to electronic analogue computers in general. The author presumably had a specific machine in mind when he stressed on page 20 that "slowly varying outputs of integrators are invariably associated with low input voltages."

The second half of the book is much better than the first and it presents an interesting review of techniques built up over a period of years in the U.S.A. both by the users and by the manufacturers of electronic analogue computers. Chapters VI and VII, in particular, collect together much useful "know-how" on analytic function generation and the representation of non-linear phenomena. The final chapter indicates a possible future trend by discussing the digital integrating differential analyser.

While regretting the shortcomings of this book it can be said of it that it does present, in easily read form, much of the "art" of using an electronic analogue computer.

J. J. GAIT.

The Cathode Ray Oscilloscope Circuitry and Practical Applications

By C. J. Czech. 340 pp. 406 figs. Demy 8vo. Philips Technical Library, Holland. Cleaver-Hume Press Ltd. London. 1957. Price 57s. 6d.

THIS is another valuable addition to the Philips' Technical Library. "An Introduction to Newcomers" is an accurate description given by the author, but he should also have added—and a reference book for the engineer. A wealth of illustrations and detail makes certain that student and professional engineers alike, just cannot fail to enjoy the lucid exposition.

Part I outlines the basic theoretical and practical limitations governing the design of the c.r.t., time-base generator, and deflexion amplifiers.

Part II covers measuring techniques so well that possible uses are converted to probables. The display of hysteresis loops, phase méasurments, time measurements, and in particular the Lissajous and other figures for some very odd frequency comparisons, should prove invaluable for many engineers.

Part III helps the oscilloscope operator, using practical examples, to choose the best methods; and finally Part IV gives detailed designs for two oscilloscopes and a time-base expansion unit.

Some omissions cannot be overlooked. They may be explained by the author's obvious determinations to choose for his examples, circuits from the products of the one manufacturer—N.V. Philips (Glocilampenfabrieken. Eindhoven).

A fuller translation for the Miller integrater circuit[1] is needed, divorced from the transitron.

D.C. coupled deflexion amplifiers should be given the space which their importance deserves.

Excepting the footnote on page 197, there appears no reference to the signal gating and waveform expansion facilities, together with the d.c. shift control and measuring techniques offered by a "long tailed" pair of valves so ably developed by the late A.D. Blumlein in 1938[2].

G. E. Luton, A.I.L., has made his valuable contribution by producing the almost flawless translation from the German. However, if he has a "Local," he should determine customers' reaction to his use of the term "opening time," when he means time open. ("between-lens" shutters, page 260).

The reviewer also noticed the following errors:—

Page 34. Confusion in the last few lines exists between the average potential between the plates, and the average potential between plates and c.r.t. cathode.

Page 50, "Synchronization of time-base circuits." In the third line of this paragraph, the word "multiple" is used in place of sub-multiple.

Page 125/126. Apparently an error in Fig. 5.39 shows the triode anode coupled to its control grid by a 15pF capacitor. This gives negative feedback in place of the positive feedback called for.

Page 168. 1 000kc/s appears as a misprint, where 1 000c/s is intended.

Page 174. In the last paragraph an additional number (18) appears in the references. The error begins with (15) in the sixth line from the end, so making (16) (17) and (18) wrong in the following lines.

Page 328. Section 5. Deflexion Amplifier. Two reference number 7's appear and number 6 is omitted.

H. L. MANSFORD.

REFERENCES

1 A. D. BLUMLEIN. British Patent Specification No 580,527.
2 H. L. MANSFORD. The Waveform Monitor. *Electronic Engng.*, 19, 272 (1947).

Radio, Television and Basic Electronics

By R. L. Oldfield. 342 pp. 271 figs. Demy 8vo. The Technical Press Ltd. 1957. Price 42s.

ALTHOUGH this book is yet another " learn all about the entire field of electronics without previous knowledge of the subject and without mathematics" book which follows the seemingly endless procession of such books from the United States, it is quite a good example of its kind and should enable the student to learn something about the fundamentals of the craft and something of the application of those fundamentals.

It is obviously impossible thoroughly to cover such a vast field in such a book and the author has been quite successful in choosing the material so that enough of the various branches of electronics are covered to give the reader a reasonable " taster ".

After a short history of electronics (which started in 1883 with the discovery of the Edison effect by Thomas A. Edison according to the author) there follows chapters on the principles of electricity, magnetism, resistance and resistors, capacitance and capacitors, inductance and inductors, units and Ohms law. Direct and alternating currents and measuring instruments are next dealt with and these chapters are followed by descriptions of valves and other electron devices. Transmitters and receivers are then dealt with including associated components such as loudspeakers and aerials.

Television (which "was barely out of the development stage when it was introduced commercially in 1946" . . . . "met with instantaneous and enthusiastic reception with the American public. . . .") is then briefly described, and the book ends with a short chapter on transistors and a glossary of electronic terms.

The book is plentifully illustrated and at the end of each chapter there is a list of review questions. Unfortunately, however, things may be read which are not particularly helpful to a student. On page 13, for instance, when describing the action of a simple voltaic cell, the author states ". . . . in other words, the flow of electricity is now known to be the electron flow, which is exactly opposite to the previous theory of current conduction or current flow. In order to prevent confusion, however, the conventional current flow will be used. . . . ".

Apart from the usual difference of American terms and spelling others appear in this book. Sulfer, sulferic acid, zinc sulfate, are examples.

However, only a few errors were noticed, page 152 for instance shows an incorrect circuit diagram of a full wave rectifier circuit.

This, then, may be the book for a student wishing to know something about the wide field of electronics without going into the various branches at all deeply.

C. H. BANTHORPE

Mathematics for Electronics with Applications

By H. M. Nodelman and F. W. Smith. 391 pp. 125 figs. Medium 8vo. McGraw-Hill Book Co. New York and London. 1956. Price 52s. 6d.

THE aim of the authors of this new book is to lead the reader across the barrier which, according to their teaching experience, often exists between mathematics and its application. The book is based on a collection of engineering problems which have been taken directly from the leading periodicals in the electronics field and which serve to demonstrate the practical uses that industry makes of mathematics. With the minimum of mathematical theory the authors set out to show how determinants, matrices, series, and differential equations enter into the study of electronic systems.

The first four chapters of the book deal with the application of elementary calculus to some simple electronic problems and the use of dimensional analysis in checking the validity of equations. These are followed by four chapters devoted to the application of determinants and matrices to electronic network analysis, but the treatment is marred by errors and misleading statements. In the introduction to determinantal analysis a theorem is given concerning the sum of two determinants which is obviously false. The discussion on network analysis contains the statement that the nodal method cannot be used without modification when mutual induction is present because the voltage across an inductor coupled to another inductor is dependent upon two different currents. The essence of nodal analysis consists in the point of view that the current in an inductor coupled to another inductor is dependent upon two different voltages. In the section on network topology it is stated that a network which can be drawn on a plane, but which cannot be drawn on a sphere without intersecting branches contains two separate parts. In fact, if a network can be drawn on a plane without intersecting branches it can be so drawn on a sphere. This property defines a planar network and has nothing to do with the number of separate parts.

The next two chapters are concerned with series and their uses in the solution of engineering problems. Special reference is made to power-series representations of systems with non-linear characteristics. Two chapters on differential equations follow in which the derivative operator and the method of the Laplace transform are introduced and the book concludes with a brief account of the principles of Boolean algebra and a mathematics study plan for electronic specialists which includes an analysis of mathematical usage in the technical literature.

There can be little doubt that the need exists for a book such as this purports to be, but one is left with the feeling that an opportunity has been missed. If the discussion of differential equations and the Laplace transform had preceded that of matrices the real significance of the latter in network analysis could have been made evident. The combination of the Laplace transform and the algebra of matrices constitutes an efficient implement for the study of linear systems. By adopting such an approach the harmonic steady-state response could have been allowed to emerge as a particular case and the time and frequency domains properly correlated. The chapter on determinantal analysis could have been omitted had the treatment of matrix analysis not been confined to the 2-port theory of Strecker and Feldtkeller. Presented along these lines the book would have served more effectively to put the electronic engineering student in touch with the mathematical techniques he will encounter when he enters the engineering world.

S. D. DEARDS

Guided Weapons

By Eric Burgess. 247 pp. 50 figs. Demy 8vo. Chapman & Hall Ltd. 1957. Price 25s.

THIS book outlines the difficulties, technical and otherwise, which have delayed production of guided weapons. It explains the various types of propulsive systems and their limitations, the uses and forms of guidance, testing and historical background.

Each chapter is fully referenced to specialized literature which together with the general bibliography should prove of use to engineering and other students aspiring to a career in the guided weapons industry.

ELECTRONIC EQUIPMENT

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

GAMMA SWITCH

(Illustrated below)

Isotope Developments Ltd., Beenham Grange, Aldermaston Wharf, Reading, Berkshire.

THE Gamma Switch is a development in the application of nucleonics to the control of industrial processes and is an inexpensive and general purpose industrial device based on radiation detection.

Level control, density and liquid interface detection, and flow failure detection are examples of the type of control which can be achieved very simply and reliably using the switch. The following are among the advantages of the technique, compared with non-nucleonic methods:

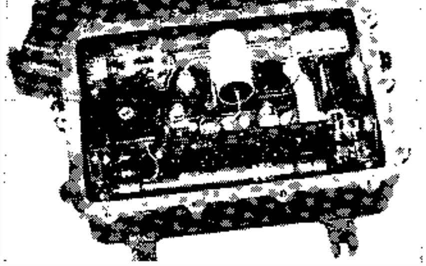
(1) No probes, floats, electrodes, or other electrical or mechanical connexions susceptible to damage are needed inside the hopper, tank, pipe, etc. The device is mounted completely externally.

(2) It is not necessary to put holes through the vessel to which the switch is attached. When appropriate it can be mounted independently to avoid the ill-effects of heat, vibration, etc., transferred from the vessel.

As the name indicates, it is a switch rather than a precision measuring instrument. It consists of a halogen quenched Geiger-Muller tube in circuit with a simple relay, built on to a printed circuit chassis and contained in a fireproof or weatherproof cast alloy case of approximate dimensions 11in by 10in by 4in.

It connects to a normal mains supply, and its relay contacts can be wired direct into conventional control circuits. The relay operates when a predetermined intensity of gamma radiation is detected by the Geiger-Muller tube.

As the switch only operates when the radiation received increases or decreases beyond a predetermined level, it gives no indication of changes in radiation above or below this level. This is why it is inexpensive compared with other nucleonic instruments which give a continuous indication of the amount of radiation received and can therefore be used for precision measurement.



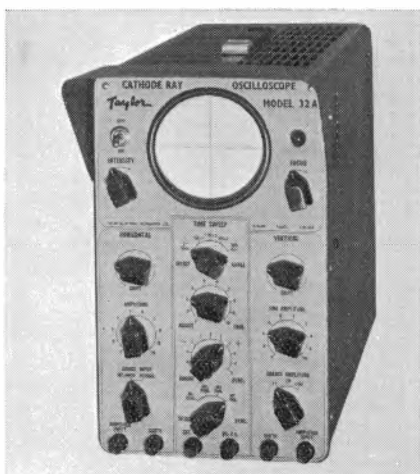
PORTABLE OSCILLOSCOPE

(Illustrated below)

Taylor Electrical Instruments Ltd., 419-424 Montrose Avenue, Slough, Buckinghamshire.

THIS oscilloscope is of a particularly robust nature and has been designed primarily to meet the requirements of television and radio service work, but its versatile features and advanced performance make it also suitable for use as a general purpose instrument.

A flat faced cathode-ray tube 4in diameter is utilized. A new type of graticule is used allowing ready measurements in both axes. A hard time-base having good synchronization characteristics is provided covering the wide range of frequencies from 2c/s to 100kc/s and can be operated either free running or triggered. Horizontal and vertical



amplifiers with push-pull output are provided. The latter is of high gain and has a frequency range extending from a few cycles per second to 6Mc/s while both amplifiers have outputs corresponding to several screen diameters.

These features make it readily possible to examine all kinds of waveforms, including pulse types of short duration and also to extend them for detailed examination.

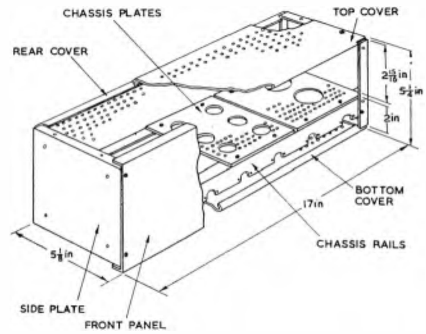
UNIT CHASSIS CONSTRUCTION SYSTEM

(Illustrated above right)

All Power Transformers Ltd., Chertsey Road, Byfleet, Surrey.

'LEKTROKIT' is a construction system consisting of a few simple and inexpensive prefabricated chassis components, from which experimental electronic chassis can be constructed quickly and cheaply.

The dimensions of 'Lektrokit' components have been designed to be compatible with a wide range of standard commercial and inter-service electronic components. Complete chassis can be assembled from a few basic parts and



the standard components available in every laboratory.

'Lektrokit' is designed to allow for almost unlimited expansion during the course of circuit development. Starting with one small chassis. the circuit can be expanded continually by the addition of further chassis until some hundreds of valve stages have been added—all without loss of neatness or accessibility.

The primary purpose of 'Lektrokit' is to save time and cost in the initial stages of circuit development, but it can also be used for more sophisticated construction, such as factory test equipment. pre-production models and even small quantity production. The addition of front panel and covers from the standard 'Lektrokit' range will convert a 'Lektrokit' chassis into an elegant and finished piece of permanent equipment.

The basic feature of the system is the unit chassis plate, a ready-to-use sub-chassis prefabricated in light alloy. Unit chassis plates are available in the following three basic patterns:—

(1) With a uniform plain pattern of perforations, suitable for mounting miscellaneous components, tag strips, terminal blocks, etc., and for wiring through from under surface to top surface.

(2) Similar to (1) above, but with the addition of six multi-purpose punched hole groups for valve holders. These punched groups will accept, interchangeable, standard valve holders for B7G or B8A or B9A valves.

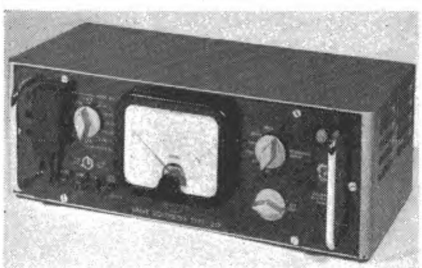
(3) Similar to (1) above, but with the addition of two large punched holes, suitable for Octal valve holders or other valve holders of equivalent size.

These three designs of unit chassis plates suffice for the assembly of a very wide range of circuit arrangements. using standard components currently available. Further designs are in progress, however, to allow for the mounting of more specialized components and further patterns will be introduced from time to time to keep pace with the introduction of new standard components.

Unit chassis plates are supported and grouped by mounting them on a pair of chassis rails. a basic working unit consisting of four chassis plates. The basic working unit thus provides a chassis area of 17in by 5in and will accommodate up to 24 valves. The construction is such that this high valve-density can be used without excessive temperature rise.

The final support for the two chassis rails of a basic working unit is provided by the addition of two standard end plates. The end plates are finished in grey Hammertone enamel, giving a completely finished appearance to the assembly.

The standard end plates have holes to allow for the fitting of front panel ventilated covers and rack mounting brackets, all of which are standard items in the 'Lektrokit' system. These holes may also be used to bolt two or more end plates together thus allowing the chassis to be expanded both vertically and horizontally to almost any desired extent.



VALVE-VOLTMETER

(Illustrated above)

Airmec Ltd., High Wycombe, Buckinghamshire.

THE valve-voltmeter type 217 is an accurate general purpose instrument suitable for measuring balanced, unbalanced and differential a.c. voltages in the frequency range 10c/s to 200Mc/s, positive and negative direct voltages up to 500V, and resistances from 100Ω to 100MΩ.

The instrument is a new and improved version of the Airmec valve-voltmeter type 712. The re-design has resulted in a wider frequency response, higher stability, stronger chassis construction, improved probe design, simpler probe mounting and easier lead stowage.

Voltage measurements in the frequency range 10c/s to 5Mc/s may be made with the probe mounted by employing the terminals provided on the front panel. These terminals, which are only brought into circuit when the probe is mounted, have large diode capacitors permanently connected across them for maintaining the low frequency response. Above a frequency of 1kc/s measurements are normally made with the low capacitance h.f. probe which when unplugged from the front panel, can be used up to a distance of 3ft away from the instrument. On replacing the probe the lead simply slides back out of the way into a special compartment inside the instrument. It will be noted that at frequencies between 1kc/s and 5Mc/s either the front panel terminals or the probe may be employed.

Positive and negative direct voltages are measured by using the terminals on the front panel in conjunction with the polarity reversing switch, the input resistance under these conditions being approximately 40MΩ.

Six resistance ranges are provided giving readings up to 100MΩ. The circuit employed is such that zero ohms is indicated by zero meter deflexion and infinite resistance by full scale deflexion. The maximum test voltage that can appear across the resistor being measured is 5V when the meter reads f.s.d. (i.e. infinite resistance), and this decreases in proportion to the meter reading.

The completely balanced circuits first introduced in their model 712 have been retained, and the stability has been further improved by the use of fully aged and matched modern type valves. As a result the meter zero does not change by more than 2 per cent of full scale deflexion on the lowest d.c. range for a 10 per cent change in mains voltage, and on higher ranges the zero change is negligible. Long term zero drift is equally small.

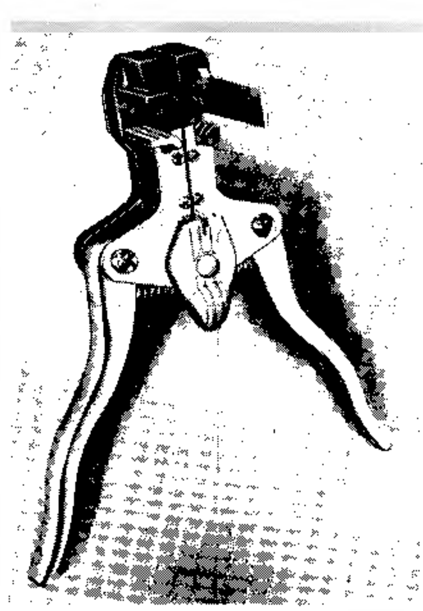
WIRE STRIPPERS

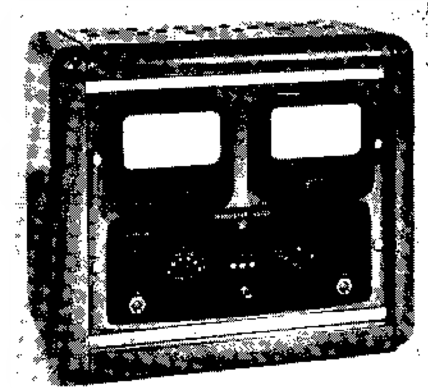
(Illustrated below)

Erma Ltd., Hong Kong Works, Exhibition Grounds, Wembley, Middlesex.

THE 'Trig-O-Matic' is a new model wire stripper embodying a patented trigger mechanism which keeps the jaw bodies apart until the stripped wire has been withdrawn. This feature prevents the buckling of wire strands during the stripping operation and is specially valuable when dealing with finely stranded cable. In appearance this tool is similar to the existing 'Standard' model.

Both tools accommodate a cable range from 0·014in to 0·144in conductor diameter, using interchangeable cutting blades of overlapping sizes. The wire strippers may also be used for cutting wire and a stripping gauge is fitted.





TRANSISTOR TEST SET

(Illustrated above)

Bobochord Ltd., Lancaster Road, High Wycombe, Buckinghamshire.

THIS instrument has been designed to a customer's requirements but it is now available for general release. It has been designed to test such transistors as are used in low power audio devices but it can be rendered suitable for testing high power transistors in an alternative form. The instrument has been designed with the following specific tests in mind and it offers advantages not generally possessed by competitive instruments.

Measurement of I_{co}.

In the grounded emitter configuration, the collector current with zero base current, I_{co}, can be measured for any chosen value of junction voltage, V_c.

Measurement of α', the short-circuit current gain in the grounded emitter configuration.

This is an actual a.c. small signal measurement using a 10μA r.m.s. input base current of frequency 1kc/s.

Initially, the 10μA base current is set up accurately by adjusting the CAL control, and then by switching to the α' position a direct reading of α' is automatically obtained.

This measurement can be made at any junction voltage V_c. The collector current, I_c, can be varied continuously and is permanently displayed, thus enabling its effect on α' to be readily observed. This facility enables the optimum collector current, I_c, to be measured for any individual transistor.

Noise.

Since transistor noise is of importance in high gain transistor amplifiers, a facility is provided for the quantitative assessment of the noise generated by a transistor. The transistor noise factor can be obtained for any combination of junction voltage, V_c, and collector current, I_c.

NEW INDUSTRIAL VALVES

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

THREE triode valves, intended primarily for use in industrial heating equipment, have been developed. With thoriated-tungsten filaments and anode dissipations of 5, 10 and 20kW, the forced air-cooled versions are the ACT100, ACT101 and ACT102 respectively. Water-cooled versions of the 10 and 20kW types are also available.

Short News Items

The Society of Instrument Technology is setting up a special section to cater for the current interest in data processing systems. This section will begin to hold meetings in London this autumn and four papers are scheduled for the first session. 10 October: "A System for Handling Wind Tunnel Data", by J. F. M. Scholes of R.A.E. Bedford. 14 November: "A Digital Plotting Table", by J. Morrison of Dobbie, McInnes Ltd. 28 January 1958: "Digital Codes and Coding", by M. P. Atkinson of N.P.L. 29 April 1958: Joint meeting on "Scanning and Logging", with papers by D. H. Whiting and J. Dunkley of I.C.I. and J. Churchill of Sunvic Controls Ltd. Further information may be obtained from the Secretary, S.I.T., 20 Queen Anne Street, London, W.1.

The Society of Industrial Radiology will now be known as The Non-Destructive Testing Society of Great Britain. The address remains at 2 Tomswood Terrace, Barkingside, Essex.

With the complete integration of Siemens Bros & Co Ltd and The Edison Swan Electric Co Ltd, the company will no longer be distributing Tungsram valves. Orders will be taken over by the British Tungsram Radio Works Co Ltd, West Road, Tottenham, London, N.17. Siemens Edison Swan Ltd will be responsible for the manufacture and distribution of all Ediswan and Ediswan Mazda valves and cathode-ray tubes.

P.A.M. Ltd, of Merrow, Guildford, have appointed Robshaw Bros Ltd of Bournemouth as sole distributors of Nera projection television receivers for the southern counties.

The BBC have announced that a new v.h.f. sound broadcasting station for North-East Wales will be built on a site 1 800ft above sea level at Cyrn-y-Brain, near Llangollen, and some eight miles west of Wrexham. The station will be known as Llangollen. Building work will start as soon as possible, and it is expected that the station will be completed by the autumn of next year.

Standard Telephones and Cables Ltd have received an order from the Royal Rhodesian Air Force for a complete Ground Controlled Approach System for one of their principal military airfields.

Pye Ltd of Cambridge recently demonstrated the first battery-operated industrial television system on Swiss railways. In Geneva the equipment, which operated off two 12V car batteries in series, was demonstrated viewing rail alignment from a moving train. Railway engineers were thus able to watch the track on a receiver inside the train. This equipment can also be used for examining springs and other inaccessible parts while the train is in motion.

Pye Ltd also announce that they have sold a complete high-power television station for Bangkok for £170 000. It is expected to be completed towards the end of this year. In 1955 Pye sold a complete television station for Baghdad to the Iraq Government.

Klockner Moeller England Ltd have moved to new premises at 7 Charterhouse Buildings, Goswell Road, London, E.C.1.

The Automatic Coil Winder and Electrical Equipment Co Ltd have changed their name to AVO Ltd, Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

Electronic Products have now moved to premises at Lawrence House, Breakspear Road, Ruislip, Middlesex. Additional machines have been installed for all types of layer, wave and toroidal winding.

The Post Office is making arrangements for trials of mobile radio equipment using 25kc/s channelling instead of the 50kc/s channelling equipment so far authorized. In the meantime, frequency allocations for private mobile radio services are to be made only for 50kc/s channel spacing in both high and low bands and equipment used must, therefore, meet the minimum technical requirements of the 50kc/s specifications which became effective from 1 January 1957.

The Mansol Ceramics Co of Belleville, New Jersey, have opened an overseas manufacturing plant at Thornton Heath, Surrey. The new plant will manufacture glass preforms for hermetic sealing. An interchange of ideas and technical information will keep both plants operating in close co-operation.

Sunvic Controls Ltd have opened an office at 24 Northumberland Road, Newcastle upon Tyne.

Electronic Associates Inc. of Long Branch, New Jersey, have established a European Computing Centre at 43 rue de la Science, Brussels. It is a pioneer effort to bring analogue computer benefits within the financial reach of all Western-European industry and business.

The Wayne Kerr Laboratories Ltd announce the appointment of Mr. R. Richards as Technical Sales Representative. Mr. Richards has been associated with the industry for many years and was formerly Senior Technical Sales Representative, Southern Area, for Marconi Instruments.

Radio Heaters Ltd have opened a new research laboratory at Wokingham, Berkshire, to do work on h.f. induction and dielectric heating equipments of all types. The services of this new laboratory will be available to any potential industrial user of this type of equipment.

The Plessey Co Ltd announce that production of their standard communications equipment for industrial, commercial and Service applications has been moved from the Ilford, Essex, factory to the St. Ives Unit, West End Mills, Ramsey Road, St. Ives, Huntingdonshire.

H. M. Hobson Ltd of Wolverhampton have recently purchased an Elliott G-PAC analogue computer to aid them in their advanced studies relating to the dynamic performance of fuel systems and servo control units.

Mr. Leonard Bennett has been appointed Technical Secretary of the Radio and Electronic Component Manufacturers' Federation.

Mr. D. C. Rogerson has been appointed Deputy Publicity Manager of The General Electric Co Ltd, responsible for the company's publicity on heavy engineering and atomic energy. This is an additional appointment to that of Mr. Arthur Clarkson who continues as Deputy Publicity Manager, an appointment which he has held since 1938.