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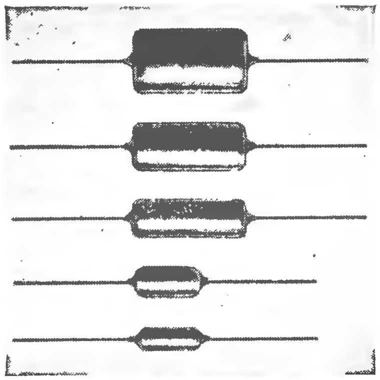
POLYESTER FOIL CAPACITORS

for improved performance and greater reliability

One of the latest developments in electronics components for television and radio receivers is the Mullard polyester capacitor. These capacitors consist of a non-inductive winding of polyester and aluminium foils which is coated with a special protective lacquer. The polyester dielectric is produced in the form of a very thin homogeneous film—as thin as 0.00025 inches—which has a higher insulation resistance and dielectric strength than paper film. Compared with the older paper types, the new capacitors possess lower dielectric losses and a higher insulation resistance, are more resistant to moisture, can withstand higher temperatures, and are slightly smaller.



Polyester capacitors are supplied in a comprehensive range of capacitances. In the 400V working voltage series, rated at 85°C, the values extend from 0.001 to 0.47μ F$\pm$10%, and in the 125V series, from 0.01 to 1μF$\pm$ 10%. With their wide coverage of values, superior electrical and physical properties, and reduced dimensions, these capacitors are making a significant contribution to the improved reliability and performance and greater compactness of today's radio and television receivers.

SILICON DIFFUSED MAINS RECTIFIER

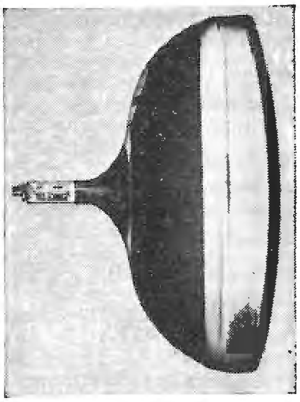
type BY100

The BY100 is the silicon diffused diode designed by Mullard as a television mains rectifier which is now to be found in many present-day receivers. It is produced by the alloy-diffusion process, which gives excellent uniformity of characteristics and ensures good reliability.

The BY100 has a maximum recurrent peak inverse voltage rating of 800V, at which the reverse current is 10μA. It will deliver an average forward current of 550mA at 50°C, while its maximum peak current rating is 5A. At this peak, the forward voltage drop is 1.5V.

The length of the rectifier is about 2cm and its diameter is about 1cm. This small size is an obvious advantage: it allows greater flexibility in positioning the device in a receiver and enables localised heating to be minimised. The BY100 thus offers convenience of use and is also contributing effectively to the very high standards of performance evident in so many modern television receivers.

MULLARD 'SHORT-NECK' PICTURE TUBE



Better electrical performance is combined with shorter overall dimensions in the new Mullard picture tube, type AW47-91. Electrostatic focusing and a unipotential focusing lens are used. Good spot quality is thus achieved over the whole picture area. An ion trap is not required, so that the spot quality is further improved. The electrode gun of the AW47-91 is very short, and the sealing pip at the base of the gun is very small. The total neck length is only 110mm, and this short-neck construction has contributed considerably to the production of modern slender television receivers. The electrical and physical advantages of the short-neck construction make the AW47-91 an important advance in picture tube design.

WHAT'S NEW IN THE NEW SETS

These articles describe the latest Mullard developments for entertainment equipment

MVE 1080

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Over to 625

SINCE its inception in 1936, the 405-line standard has been consistently supported by this journal: first as a bold and imaginative anticipation of future needs by Sir Isaac Schoenberg and his colleagues in E.M.I.; later in opposition to the establishment in Europe, *circa* 1950, of the 625-line standard on the grounds of economy of bandwidth; and more recently because we think that the marginal improvements of a 625-line picture are insufficient to justify the cost and chaos of a changeover. Until now the verdict of the viewing public, the final arbiters in these matters, has been that 12 million of them have seen fit to buy 405-line sets and take out licences. Some are not switching on for as many hours as they used to, but although we have heard many grumbles about the content of programmes and the quality of some outside broadcasts and recordings we have yet to hear of anyone who has switched off in disgust at the inadequacy of a well-lighted 405-line picture direct from the studio. The compulsive urge to change to 625 lines—fed by some sections but by no means the whole of the domestic receiver manufacturing industry, and supported by the broadcasting authorities for whom it would ultimately make the task of exchanging and transmitting programmes easier—has not come from the viewing public; they have still to see 625 lines and pass judgment in their own homes.

But the die is now cast: the Pilkington Committee has endorsed the 1960 recommendation of the Television Advisory Committee and the Government has decided that any new programmes in the u.h.f. bands and eventually all television broadcasting in the U.K. shall be on the 625-line standard (with negative modulation and f.m. sound). We and the other viewers must now accept that decision, which has been arrived at after what the Pilkington Report* clearly shows to have been a full appreciation and a fair representation of all the facts.

We believe that the British radio industry has the capacity and, given the will, could already be competing in the overseas markets for 625-line receivers, and that the broadcasting authorities could quite easily duplicate their pick-up and recording devices to serve home viewers on 405 lines and the Eurovision network on 625 (even the American continent on 525); but we find it harder to refute the conclusion that (paragraph 753 of the Pilkington Report) "The main advantage of the 625-line over the 405-line standard is that, in the long run, the public will get a better or a larger picture at little if any extra cost." There is little doubt in our mind that the public will choose the larger rather than the better picture. This is confirmed by experience in Europe where the 23-in tube size is now dominant (and where incidentally the adequacy of even 625 lines is already called in question by the adoption of line-eliminating devices of various kinds). The Committee has been careful to qualify "at little if any extra cost" by "in the long term." So we hope our children will be grateful to their parents for having spent an additional £50M to £150M (according to the method finally adopted for changeover) to make this possible.

Next in importance to the potential increase in picture size stands the *fait accompli* of international acceptance of 8-Mc/s channel spacing for new transmitters in the u.h.f. bands. Experience this summer of interference in the v.h.f. bands confirms the wisdom of this decision. With 405 lines we should be using effectively only 5Mc/s and it would be an unpardonable waste of bandwidth (the scarcest commodity in radio communication) not to use the full 8Mc/s by adopting 625 lines.

These arguments have led us to the root of the matter, namely, that we are in process primarily of extending the programme capacity of our television service. To do this effectively we must go up in frequency to Bands IV and V. There we shall find trouble enough without courting interference from foreign stations. Therefore we have agreed internationally to 8-Mc/s channel spacing and, having got it, we might as well use it. In true perspective, 625-line definition comes as a small bonus to the benefits which we hope will accrue from our courageous decision to take a few more hairs of the dog that bit us.

As a technical journal we are absolved from the duty of commenting on the merits of the Committee's proposals for the uses to which the extended services should be put, but in reading them as citizens and viewers we cannot fail to be influenced in assessing the weight to be ascribed to their social judgments by the competence and impartiality of their treatment of the technical foundations of the medium.

* "Report of the Committee on Broadcasting, 1960," Cmnd 1753, H.M. Stationery Office, Price 18s.

Classification and Retrieval of

TECHNICAL INFORMATION

1.—CONSIDERATIONS LEADING TO THE CHOICE OF NOTCHED CARDS

By A. E. CAWKELL

"*Certainly it is harde to playse everyman bycause of dyversite and chaunge of langage. For in these dayes everyman that is in ony reputacyon in the countre, wyll utter his communicacyon and matters in sych manners and termes that fewe men shall understonde theym.*"
William Caxton.
(1422-1491)

THE fund of available technical information has now reached fantastic proportions, but the effort required to extract information relevant to a problem in hand is often considered to be equally fantastic. In the words of Lord Rayleigh "re-discovery (of information) in the library may be a more difficult and uncertain process than the first discovery in the laboratory."

In most development laboratories some kind of reference system usually exists, often of a personal nature understood only by the user, who is also the abstractor and classifier.

The creation and maintenance of a system which can be used by any engineer, which is accessible, and which is of real value, is undoubtedly a formidable task. However access to the right piece of information at the right time, in terms of development effort saved, can recover the cost of the system many times over. In a recently cited case[1], after three million dollars had been spent on a cloud-seeding experiment, a librarian stumbled over a report showing that the work had already been completed elsewhere for 1/12th of the cost. Duplication of effort, albeit usually on a smaller scale than the above example, is almost certainly widespread at the present time.

It is hoped to show that it is feasible for a relatively small organization to operate an information system so long as it is carefully planned and not over-ambitious.

Criterion of Success.—The main requirements of an ideal system are as follows:—

1. All references on subjects of interest are held in the system.
2. However the searcher formulates his question it is "understood" by the system.
3. The searcher retrieves all the relevant references, without any redundant information, by a simple and rapid operation. The corollary of the first requirement is that if no references are found, none exist.

Having retrieved the reference, the work of the searcher is not at an end; his next problem is to find the document cited. This may be a report held in his library or laboratory, or maybe some book or report which must be obtained elsewhere. This operation is usually relatively simple, and although a part of the overall retrieval, will not be considered here.

Problems in Classification and Retrieval

It is assumed that we are not interested in the whole field of human endeavour, but in some small fraction of it, namely electronics and allied subjects —and only a specialized part of that; nevertheless a wide range of subjects will require to be listed. At minimum we must have a *subject index*, which in its simplest form might consist of an alphabetical list of subjects against each of which are listed references inserted there by the *abstractor*.

Let us consider that such a volume exists and that an engineer wishes to retrieve information about: "The detection of rot in timber by wave propagation measurements." He has seen an article on the subject about a year ago, and whilst remembering that it was by Smith, cannot remember the exact title or where he saw it. He refers to the subject index and finds that it is a bulky affair with loose-leaf pages, the latter having been introduced by the classifier as an aid to keeping it in alphabetical order; in spite of this, a number of pages are being revised, as initially insufficient space was left to keep lists of references against each subject in alphabetical order. The searcher now considers the most likely headings and tries "rot," "timber," and "wave propagation" without success.

He notices, in passing, a complete section under "N"—"Non-destructive testing," and does in fact find "timber" in this section, but no relevant reference. There is no author index to references; obviously the classifier is already hard put to it to maintain the subject index. After making a few rather more desperate attempts trying "trees," and even "ultrasonics," he abandons the effort, concluding that the article was never entered in the index.

Three months later he spots a familiar looking document lying in the laboratory entitled "Forestry Research Report 1960." Sure enough, the information he previously needed so badly (but which is no use to him now) is contained within its pages. As a matter of interest he again refers to the subject index and finds that the report has indeed been entered under " F." He is now finally convinced that the laboratories bibliographical index is useless, and resolves never to waste his valuable time by consulting it again.

The above may be an imaginary and over exaggerated example, but it is typical of the kind of difficulties which must be familiar to most research and development engineers and it serves to illustrate certain basic problems which must be resolved.

1. *The order of language.*

If the major or controlling noun, or the most significant word is brought to the front, the reference will appear in the most helpful alphabetical position. For instance:
" Timber, detection of rot in, by wave propagation methods."

Taking this principle further, it has been suggested[2] that to further assist the searcher, the main descriptors should be isolated and then indexed as a rotated entry in their alphabetical order, for instance:
Under ' R ' Rot—Timber—Wave propagation methods.
Under ' T ' Timber—Rot—Wave propagation methods.
Under ' W ' Wave propagation methods—Rot—Timber.

2. *Synonyms.*

It might be considered that both classifiers and searchers should be aware of the synonym difficulty; for instance if the subject title " see-saw circuit " is not found, " anode follower " would automatically be tried. Alternatively both may be listed alphabetically.

An investigation was carried out in a particular case[3], when 368 headings were analysed, which had been suggested by users as representing the subject matter of six books and which differed from those used in the official catalogue of the library. In 70% of the differences synonyms had been used, in 72% there was a different order of language and in 93% the users' concept of the subject was different from that in the catalogue.

3. *Arrangement of the index.*

What is logical to one person may not be logical to another. To the classifier " N "—" Non-destructive testing " may be an obvious classification for " detection of rot in timber," but this may not necessarily occur to the searcher. On the other hand, if the laboratory is frequently concerned with non-destructive testing, it may be helpful to have classification " Non-destructive testing," and to sub-divide that into sections, for instance, " Methods" " Materials " and " Properties," these being the facets usually encountered.

Elaborating still further, but at the risk of making the index bulky, the reference could be entered under " Non-destructive testing," and also rotated under " R " " T " and " W."

4. *Abstracting*

In only a small number of papers will a title embrace all facets of the paper. In our example it was clearly ridiculous to indicate as comprehensive a document as " Forestry Research Report 1960 " under " F " only. In this and in many other cases, it may be necessary for the abstractor to know his subject, list the title of each paper, and moreover if there is some item of considerable interest contained in any paper to which the title of the paper gives no clue, list this subject also. For instance, in our article " Rot in timber " etc., might be found a section on " The design of directional piezoelectric transducers." This may be of value to some future searcher who may not be interested in the main subject, and would be worth while appropriately indexing.

5. *Combinations of Concepts.*

When the classifier and searcher are only interested in single subjects, the index can be a relatively simple affair. The need for providing for subject combinations is however well recognized in existing systems—for example U.D.C., Colon, College of Aeronautics[4], and others. In electronics the interest may often be in a reference having a combination of subjects; an example of dealing with this by " rotated entry " has already been given. This gives rise to considerable effort and usage of index space by the classifier, but additionally there may be other concepts which it is desirable to index. In general these will not be as fundamentally important as the subject matter; the need for including them is a matter of individual consideration; some examples follow:—

(a) Author's name; this can be important particularly when an author becomes a specialized subject authority. Referring again to " timber testing," it is known that Lee is an authority, but his writings have a wide variety of titles. It might be much quicker if all references by Lee could be retrieved, should a general knowledge of the subject be required, than if a search was made through subjects or combination subjects.

For any reference, an author's name if remembered, offers a very selective means of retrieval particularly if it can be closely identified with a particular field.

(b) Scope of reference; it can be frustrating if when 20 references are given under a particular subject, 16 are of an elementary nature, and these have to be scrutinized before the searcher, who in this case is imagined to possess the elementary knowledge, arrives at the reference required. Supplementary information against each reference can be of value—for instance " advanced treatise," " paper," " fundamental article," " short note," " working circuits given," " patent specification," etc.

(c) Date of publication; if the approximate date of the reference required is known, and if in some way it is relevant to the reference, this may offer a means of selective retrieval.

In fields where development is rapid, computers and transistors for instance, it would be a waste of time having to refer to a number of references spread over the years, when the development cited in the article of interest is known by the searcher to have started in say 1958.

Limitations of the Conventional Catalogue Index

It will be realized from the foregoing that if the various facets mentioned are considered to be of sufficient importance to be embodied in the classification system, the compiling and maintenance of the index will become a very laborious task. The index so far considered is a very simple one, and although additional effort will be required, supplementary information giving, for instance, author's name, date of publication, etc., might be considered essential. If necessary a reference could be expanded into an abstract, giving the searcher a chance of rejection at the outset, instead of it being necessary for him to obtain the documents referred to. Furthermore, to make for ease of retrieval, the "rotated entry" system or some similar artifice may be necessary, in which case each complete reference or abstract will have to be entered in a number of alphabetical positions. At this stage the maintenance and bulk of the catalogue probably becomes prohibitive. It is mainly for these reasons that a number of authorities [5,6,7,8], have changed over to punched cards for classification and retrieval; certain other advantages accrue from this relatively new technique.

Punched Cards

Two ways of using punched cards for the classification and retrieval of technical information have been observed.

1. *The "subject" card system.*

This method, as exemplified by the "Peek-a-boo" system[9], consists of one punched card per subject, the subject title being written or typed at the top of the card, the cards then being filed in alphabetical order.

A rectangular space is available on each card of 100 millimetres × 200 millimetres, and a transparent graticule consisting of say 100 horizontal lines marked 0 to 100, and 200 vertical lines marked 0 to 200, can be laid over the card.

The ordinates of 20,000 points on the cards may now be identified by means of the grid.

In the "Peek-a-boo" system each point represents the number of a document, so that each card can accommodate 20,000 numbers representing 20,000 documents. If, for instance, the subject heading on one card is "amplifiers" and the subject heading on the 2nd card is "magnets, magnetism, magnetic" then the abstractor upon reading a document about "magnetic amplifiers" which might be allocated the number 4,967, would punch hole number 4,967 on both the "amplifier" and "magnetism" cards. When retrieving, a searcher interested in magnetic amplifiers will align the two cards behind the graticule, observing hole coincidence (if necessary using a light). The numbers of any documents about the combination of "magnetic" and "amplifiers" may then be obtained.

The "Uniterm"[10] system works on a similar principle except that the document numbers are written on to subject cards, and the occurrence of the same number on the subject cards of interest indicates that the numbered article is about a combination of those subjects.

2. *The "document" card system.*

In this system there is one punched card per document, any relevant information being typed or written on to the card.

The presence or absence of a hole indicates the presence or absence of a subject. In the case of a hundred subjects of interest, for instance, position number 8 might indicate "amplifiers" and position 29 "magnetic." Any card bearing information about a document concerning "magnetic amplifiers" would then have a hole punched in the two positions. A sort can then be conducted to retrieve only those cards about "magnetic amplifiers."

Both methods have their place; for the purpose under discussion "document" cards are considered to be more suitable for the following reasons:

(a) Initial cost of hand-sorted punched card systems is lower.
(b) Serial numbering of documents or references is unnecessary.
(c) Order of cards in file is immaterial.
(d) Abstract, author, and other information is immediately obtainable by reference to the card.
(e) Cards may be fed into the system by a number of abstractors, working at different places.

Hand-sorted "Document" Cards

It is considered that the well-known machine-sorted punched cards, requiring expensive associated equipment, are outside the resources of an average development division. The purchase of 10,000 hand-sorted punched cards, printed to order, with associated equipment costs about £120 and it is to considerations of a system embodying these cards, that the remainder of this paper will be concerned.

The cards are normally 8in. × 5in., and are supplied with $\frac{1}{8}$in. diameter holes punched round their edges, each card carrying about 96 holes.

The centre of the card may contain the title and author of the article, a short abstract, and details about location of the article, inserted by the abstractor. Information is supplied printed on each card indicating the meaning of each hole. A meaning will be printed opposite, the same hole on every card—for instance "magnetic amplifiers"—but more usually, because there are insufficient holes for direct meanings, some form of coding is used. The classifier, having written or typed in the abstract, considers the various facets of interest, and with a clipping tool makes the required hole into a slot round the edge of the card. The cards are kept in a drawer in any order, and have one clipped off corner to ensure that they are filed the right way up.

A searcher interested in say, "magnetic amplifiers," then inserts a needle through the pack at the designated hole, and all those cards which have been slotted will drop when the pack is lifted. The dropped cards will carry the abstracts of interest.

Often one or more holes out of a group of holes are slotted to indicate a meaning. This group of holes is known as a "field."

There follows an example illustrating the fundamental difference between indexed catalogues and punched cards (after Ball).[11]

Suppose cars are required to be classified in a catalogue; for stock records.

Chevrolets	New	8-cylinder		
			Blue—	Nil.
			Grey—	Nil.
			Yellow	1 in stock.
		6-cylinder	Blue—	Nil.
			Grey—	Nil.
			Yellow	4 in stock.
	Used	8-cylinder		
			Blue—	Nil.
			Grey—	Nil.
			Yellow	2 in stock.
		6-cylinder		
			Blue—	Nil.
			Grey—	Nil.
			Yellow—	Nil.

For Fords, Plymouths, Buicks, etc. the above table would require to be repeated in its entirety.

On cards there would be four independent fields, and a sort could rapidly be carried out using four needles at once to select those cards on which the desired characteristics are satisfied in each field:—for instance, Chevrolets, used, 8-Cylinder, yellow.

1.	(make)	Chevrolet √
		Ford
		Plymouth
		Buick
2.	(condition)	New
		Used √
3.	(engine)	8-cylinder √
		6-cylinder
4.	(colour)	Blue
		Grey
		Yellow √

One card would drop out showing the stock position of the car with the desired characteristics.

It will now be evident that a dimension has been added which is not present in the catalogue. If it is desired to take full advantage of this dimension, references can be simultaneously grouped according to author, subject, application, format, publication, date, and a host of other details any combination of which may be retrieved at will; even combinations not considered during classification can be retrieved.

Enthusiasm must, of course, be tempered with the knowledge that only details of real interest should be included, as otherwise unnecessary multi-aspect classification will follow, causing running costs to soar.

The Index when Using Cards.—It may seem strange that it is necessary to have an index to files of punched cards; the purpose of the index is as follows:—

1. To identify the punching code for any subject (such a code is necessary unless direct subject coding is used).
2. To establish a hierarchy of things, functions, processes, etc. in alphabetical order, showing the code number (should the latter be necessary).
3. To establish a standard terminology understood as well by the searcher as by the classifier, and to control the terms and relations used in the system.

Arrangement of the Index and Retrieval.—The mechanism of retrieval will be strongly influenced by the arrangement of the index. The view may be taken that all material may be divided into a comparatively few major subjects, which we shall call "A" level subjects. These can be subdivided into "B" level groups; these in turn can be further subdivided, and so on perhaps down to "D" level; the subject becomes more specific at each division.

For example "amplifiers" might be an "A" level subject divided into a number of "B" level subjects say "DC," "Wideband," "Audio," etc., each in turn divided into "C" levels.

As an example of each level consider

"A" level;	Amplifiers.
"B" level;	Audio.
"C" level;	Microphones.

The methods of retrieval will be to initially needle "Amplifiers", upon which say 500 cards will drop; then needle the 500 for "audio", producing say 100 cards, then needle the hundred for "microphones", giving say 5 cards. Consider the ideal case of a system of equal distribution of cards at each level; out of 10,000 cards imagine ten "A" level subjects with a thousand cards each, subdivided into a hundred "B" level subjects, with a hundred cards, a thousand "C" level subjects with 10 cards, etc. Four operations only are now required to isolate one card. The first sort eliminates nine-tenths of the pack, the second sort nine-tenths of the nine-tenths and so on, so that the fourth sort leaves:—

$$0.1 \times 0.1 \times 0.1 \times 0.1 \text{ of } 10{,}000 = 1 \text{ card.}$$

A mass of irrelevant detail is eliminated by a narrowing down process. To be effective the "A" level headings must be so chosen that the abstractor and searcher, when confronted with a subject, will have no difficulty in deciding under which heading the subject belongs.

However, a positioning which is logical to one individual may be illogical to another, particularly in marginal cases which are likely to be frequent. To overcome this difficulty it may be considered desirable to list all headings in strictly alphabetical order of most significant word.

Coding for Punched Cards

The necessity for adopting a code arises because there are usually a large number of concepts, but only a limited number of punched holes in a card

Direct Codes.—A direct code uses a single hole to denote a single subject or meaning; this may be printed on all cards. For a limited number of concepts this method is simple and convenient.

Combination Codes.—By a combination code is meant one in which one or more holes are slotted to represent a meaning. Many more code numbers (representing subjects or concepts) can be entered on a card, than with a direct code. The selector, sequence, superimposed, and other codes to be described are all combination codes.

Selector Codes.—To understand this code and other codes to be described later, it is necessary to consider the number of different meanings that can be signified by combining a given number (i.e. a "field") of holes, in different ways.

If C is the number of combinations of H things taken Y at a time (mathematical shorthand Hc_y), then

$$C = \frac{H!}{Y!\,(H-Y)!} \qquad (1)$$

where C = number of concepts that can be accommodated in the field.

Y = number of symbols (and hence holes slotted) in the code.

H = number of holes in the field available to be slotted.

For instance using a 2 digit code in a field of 5 holes.

$$C = \frac{5\times4\times3\times2\times1}{2\times1\,(3\times2\times1)} = 10$$

It can also be shown that C_{max}, the maximum number of concepts that can be coded is when

$$Y = \frac{H}{2}$$

Thus when slotting combinations of three holes in a field of six holes, 20 concepts could be accommodated.

The selector code shown in Fig. 1 is an example of this principle. One field is used per decade.

An important feature of any code is the number of

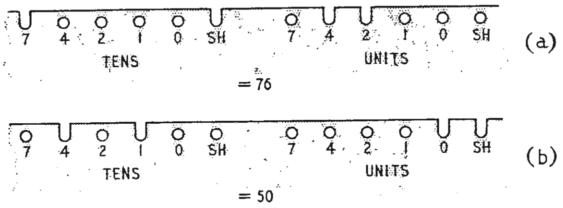


Fig. 1. Examples of selector code.

needles required when retrieving—inconvenience increases with the number of needles.

In our examples two needles are always used per decade—the hole " SH " (single hole) is used when only one number is slotted in the field. The purpose of the " SH " hole can be appreciated by reference to Fig. 1(a). If " SH " were not slotted in the " tens " field, any cards carrying 8 (7 and 1 slotted) and 9 (7 and 2 slotted) as " tens " would also drop when needling for 7.

It will be noticed that some efficiency has been sacrificed in the interests of using the easily understood decimal notation; if binary coding were used, many more concepts could be coded into the same number of holes. The system is restricted to accommodating one concept per field—slotting another number would allow unwanted cards to drop. Thus slotting into Fig. 1(b) would allow any cards coded 30 and 60 to drop.

Sequence Codes.—Sequence codes make more efficient use of holes than do selector codes, and also enable cards to be placed in a pre-determined sequence with less effort.

Values of Y are permitted to vary from zero to H.

$$C = \sum_{Y=0}^{Y=H} \frac{H!}{Y!\,(H-Y)!} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

Unlike the selector code, the number of slots per field, and hence the number of needles for retrieval, varies.

A common sequence code is 7, 4, 2, 1. For " O " no slot is made, for " 1 ", one slot, for " 3 " (2+1), two slots. When needling for cards coded, say 70, cards carrying 71 to 79 will also drop, so that this code will often be unsuitable for selection. Thus when using tens and units fields, " 70 " would be coded with only one slot—" 7 " in the tens column, whilst " 63 " would require four slots—4+2 and 1+2. As with selector codes, only one concept can be coded per field, otherwise unwanted cards will drop when sorting.

For sequence sorting any number of cards randomly arranged can be rearranged in sequence with the minimum number of operations as follows:—

> Needle the right-hand hole; place the drop outs at the back of the pack. Needle all cards in the next hole from the right; place the drop outs at the back, etc., etc.

The number of sorts necessary to serially arrange the cards is the same as the number of holes in the field.

A remarkable consequence of this phenomena has been pointed out by J. C. Kendrew[12], and is mentioned here for general interest, whilst probably not being of interest for information retrieval. This is the fact that (2^n-1) cards can be sorted into perfect serial order by n successive operations. Thus in the 7, 4, 2, 1 code already sited (2^4-1) cards—15—can be serially sorted in four operations. This method of sorting becomes remarkable when huge numbers of cards are used.

For example, if the half-million entries in the *Concise Oxford Dictionary* were each coded on a card (half a million $=2^{19}$ approximately), and randomly arranged, only 19 operations would be required to get all cards in exact alphabetical sequence although admittedly the mechanics would be difficult.

Random Superimposed Coding.—With this method of coding a much larger number of concepts can be coded into a given number of holes.

If there are 1,000 or more subjects in the index requiring a code allocation, a four-symbol code must be used. Of the various methods available, it has been shown that the most efficient is to slot the code symbols into one large field 13, as in Fig. 2, where 4 groups each containing four symbols taken from a random number table have been slotted into a 30-hole field.

The implications of this method are far reaching, and space does not permit of a full explanation, for which the reader is referred to the literature.[13, 14, 15]

The first point which is evident from Fig. 2 is that 13 slots have been used to code 16 symbols because of overlapping. To find the average number of slots required in any particular case use

$$G = H - H\,\frac{(H-Y)^X}{H} \quad \ldots\ldots\ldots\ldots \quad (3)$$

where G = number of holes actually slotted.
H = number of holes in field.
Y = number of symbols in code.
X = number of concepts simultaneously entered
and XY = max. number of slots made in the field.

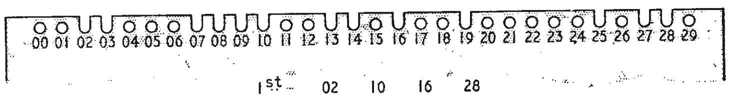
For a 30-hole field, a four-symbol code, and 4, 3, 2, or 1 concepts, G becomes 13, 10.3, 7.5, and 4, where XY is 16, 12, 8, and 4 respectively.

It is also evident from Fig. 2 that not only the cards carrying the four groups shown will drop, but also those carrying " synthetic " groups, any of which could be the group for an entirely different subject.

For instance cards carrying 02, 16, 19, 25 would drop.

It is necessary to consider how many of these unwanted cards there will be as they will have to be hand sorted from the desired cards, and too many could prove an embarrassment.

A 30-hole field has been chosen because this



1st	02	10	16	28
2nd	27	16	08	07
3rd	09	25	19	07
4th	14	03	13	25

Fig. 2. Four 4-symbol random number groups coded in a 30-hole field.

represents the number of usable holes along one edge of a standard 8×5 inch card.

If needles are inserted at random into a pack of cards already slotted with random numbers, it can be shown that the average fraction of the pack (F_d) which will drop is:—

$$F_d = \frac{\dfrac{G!}{Y!(G-Y)!}}{\dfrac{H!}{Y!(H-Y)!}} \quad \ldots\ldots\ldots\ldots \quad (4)$$

The numerator and denominator can always be simplified into an expression of the form "xCy," meaning "the highest y terms in x"

Thus for G = 7, Y = 4, H = 30

$$F_d = \frac{7 \times 6 \times 5 \times 4 \times 3 \times 2 \times 1}{(4 \times 3 \times 2 \times 1)(3 \times 2 \times 1)} \times \frac{4 \times 3 \times 2 \times 1\,(26 \times 25 \ldots \times 1)}{(30 \times 29 \times 28 \times 27 \times 26 \ldots \times 1)}$$

$$= \frac{7 \times 6 \times 5 \times 4}{30 \times 29 \times 28 \times 27}$$

which is expressed as

$$\frac{7\,C\,4}{30C4}$$

The Table I has been compiled to show the number of cards which will drop from a pack of 10,000 when needles are inserted at random, for various conditions:—

TABLE I

Line No.	No. of subjects coded per card	Average no. of holes slotted per card (G) approx.	No. of subjects it is desired to retrieve (X)	Average no. of needles required to effect retrieval	Dropping fraction (F_d) (approx.)	No. of cards dropping out of a pack of 10,000
1	4	13	4	13	$\frac{13C_{13}}{30C_{13}}$	< 1
2	4	13	3	10	$\frac{13C_{10}}{30C_{10}}$	< 1
3	4	13	2	7	$\frac{13C_7}{30C_7}$	8
4	4	13	1	4	$\frac{13C_4}{30C_4}$	260
5	3	10	3	10	$\frac{10C_{10}}{30C_{10}}$	< 1
6	3	10	2	7	$\frac{10C_7}{30C_7}$	< 1
7	3	10	1	4	$\frac{10C_4}{30C_4}$	77
8	2	7	2	7	$\frac{7C_7}{30C_7}$	< 1
9	2	7	1	4	$\frac{7C_4}{30C_4}$	13
10	1	4	1	4	$\frac{4C_4}{30C_4}$	< 1

Dropping fraction for a given number of subjects coded per card, when retrieving for a given number of subjects, using a four-symbol code.

The number of cards dropping, as shown in the last column, arises when the number of needles shown in column 5 are inserted *at random.*

Consider line 7, and assume that we are going to insert 4 needles into the pack for one specific subject, where the pack consists of cards carrying abstracts for which 3 subjects were coded on each card; let us assume that 20 cards refer to this one subject. Then those 20 cards will drop plus $\frac{10C_4}{30C_4} \times 9{,}980$, which is still very nearly 77 extra (unwanted) cards. Therefore unless the number of cards for the subject in question is large relative to the number of cards shown in the last column, a large proportion of the dropped cards will be unwanted.

For any system using a 30-hole field, it will be necessary to first estimate what proportion of cards will carry 1, 2, 3 or 4 subjects respectively (in electronics more subjects than this are a rarity), and from the table consider whether the number of unwanted cards which will drop is acceptable, bearing in mind that it may be desired to retrieve on the basis of combinations of up to four subjects. The matter will again be referred to when a practical system is considered later in this paper.

Random superimposed coding methods have been developed further by Mooers[16] whereby information is stored on tape or film in the form of patterns, the method then being called "Pattern inclusion selection." Retrieval logic then becomes feasible; for example the question might be posed to the system "select the documents containing either or both subjects A and B, in all cases also containing C, but in all cases not containing D." Such methods and the accompanying machinery are however outside the scope of this paper.

(To be concluded)

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Pilkington Committee Report

ARGUMENTS AND RECOMMENDATIONS: GOVERNMENT DECISIONS AND PROPOSALS

The long-awaited Report of the Committee on Broadcasting appointed by the Postmaster-General in July 1960 was published on June 27th. The Committee was set up, with Sir Harry Pilkington as Chairman, "To consider the future of the broadcasting services in the United Kingdom, the dissemination by wire of broadcasting and other programmes, and the possibility of television for public showing; to advise on the services which should in future be provided in the United Kingdom by the B.B.C. and the I.T.A.; to recommend whether additional services should be provided by any other organization; and to propose what financial and other conditions should apply to the conduct of all these services." It received 636 memoranda from organizations and individuals which, together with other papers circulated to the Committee, total just over 850. The 342-page Report does not include the full texts of the evidence submitted to the Committee (some of this is to be published later).*

Since the publication of the Report and its 120 recommendations the Government has issued a White Paper "Broadcasting: Memorandum on the Report of the Committee on Broadcasting, 1960"† setting out its "first series of proposals" and announcing that further proposals will be put forward later in the year on those matters "which require further consideration". In this survey of the "Pilkington Report," we shall summarize the arguments put forward, the Committee's recommendations and the Government's decisions and proposals.

ORGANIZATION AND CONSTITUTION.—The B.B.C.'s fourth Charter was extended earlier this year from 10 to 12 years so that it now expires on the same date as the Television Act (July 30th, 1964) under which the Independent Television Authority was established. Now that the B.B.C. has been in existence for 35 years "it might at first sight seem that the time had come to grant the Charter for a much longer or even an indefinite term." The Committee however recommends that "The B.B.C. should remain the main instrument of broadcasting in the United Kingdom" and that the next Charter should be for 12 years (to July 1976). This is accepted in the Government's White Paper.

After reviewing various methods of financing broadcasting—by licence fees, advertisement revenue, direct payment, as with subscription TV, or by Government subvention out of public funds—the Committee states that, in principle, the licence fee system is to be preferred. An increase to £6 per annum for a combined sound and television licence (excluding any excise duty, which at present is £1 on the £3 licence) is considered by the Committee to be "not too much to pay." In support of its argument the Committee tabulates the fees paid by 12 western European countries showing that all but one other is over £5, the highest being in Austria (equivalent to £9 13s 10d) and the lowest Holland (£3 18s 10d). The B.B.C. in evidence referred to the need for adequate finance for its "research into the scientific and technical aspects of broadcasting" as "part of the public service" for which it is responsible.

The Committee recommends that "the B.B.C. should continue to be financed out of licence revenue and only out of licence revenue" and that its Charter should be amended to empower the Corporation to borrow up to £10M to obtain temporary banking accommodation and up to £20M for capital purposes.

In its White Paper the Government "accepts its responsibility to see that the B.B.C. can secure sufficient income to finance adequate services" and also accepts the recommendations regarding borrowing powers.

* Cmnd 1753, H.M.S.O. 18s.
† Cmnd 1770, H.M.S.O. 1s 3d.

Independent television's "fundamental constitutional weakness", as the Report puts it, is its "failure to reconcile its two purposes"—"to provide a service of television broadcasting which will realize as fully as possible the purposes of broadcasting and, incidentally, to provide a service to advertisers." It goes on: "As independent television is now constituted and organized, the dynamic of profitability is applied to the realization of the incidental objective—that is to the production of desirable advertising time; and the regulatory function cannot be exercised positively to ensure that nevertheless the primary and essential objective is realized." The conclusion was reached by the Committee that "so long as this is so, no system of specific controls by regulation will ensure the fullest possible realization of the purposes of broadcasting. To do this an organic change, a change of functions is required." It calls for such changes in organization and constitution to have the following effects:—

(1) To vest the reality of power in the Authority.
(2) To remove from programme planning and production the commercial incentive always to aim at maximum audiences and at maximum advertising revenue.
(3) To apply the incentive of profitability to the production of the best programmes.
(4) To promote real competition in programme production between the programme contractors.
(5) To promote competition in good broadcasting between the B.B.C. and independent television.

To this end the committee recommends the following changes:—

(a) The Authority to plan the programming.
(b) The Authority to sell advertising time.
(c) Programme companies to produce and sell to the Authority programme items for inclusion in the programme planned by the Authority.
(d) The Authority, after making provision for reserves, to pay any surplus revenue to the Exchequer.

The participation of the press in television "the two dominant media of mass communication" was investigated by the Committee at the invitation of the Postmaster-General. "The suspicion," says the Report, "of too great a concentration in too few hands of the power to influence and persuade cannot be dismissed by the argument that the power has not been used, and is not very likely to be used. . . . In no company, however independent television is constituted and organized, should the press influence be dominant [not the largest single interest]."

Referring to these recommended radical changes in the structure of independent television the Government's White Paper states "Full account will be taken of the views expressed in public debate and the Government will later submit to Parliament a statement of its own proposals for the future of independent television."

TELEVISION.—The Committee envisages a total of six national programmes—two in the v.h.f. and four in the u.h.f. bands. It does not think that the possibility of a seventh programme, which would be feasible if the 405-line standard were retained, overrides the advantages of a change to 625 lines, as recommended by the T.A.C. in their 1960 report. It is pointed out that the cost of changing the line standard will be small compared with the total expenditure necessary to provide extra programme services in the u.h.f. bands.

Alternative methods of introducing the new wavebands and changing line standards have been carefully weighed and in the opinion of the Committee the "duplication method" in which the existing v.h.f. services will be transmitted also at u.h.f. during a transition period of 10 years or more is to be preferred to the "switchover method" in which there would be a simultaneous change of line standards in the v.h.f. bands towards the end of the period. The Committee sees more flexibility in the duplication method to accommodate possible difficulties in establishing adequate national coverage with u.h.f. It also strongly recommends co-siting of transmitters, since B.B.C. and I.T.A. as public bodies have an equal duty to ensure the transmission of the best television signal.

The Government has decided that the new u.h.f. programmes shall be on 625 lines, that test transmissions shall start this year and that a public service should be started in the u.h.f. band in London in mid-1964. The planning and rate of extension will depend on experience with the new frequencies. All u.h.f. programmes for the same area will be transmitted from the same mast.

The Pilkington Committee recognizes the future desirability of a compatible service of colour but thinks that the efforts of the broadcasting authorities should first be concentrated on developing the 625-line services. The Government feels that colour must have a place in the future pattern of television and that its introduction should not be postponed indefinitely because of cost.

Chapter XXI of the Report investigates the pros and cons of subscription television and summarizes the evidence given by several companies interested in providing either the "toll" programmes or the technical means of metering such a service whether distributed by wire or by radio. The Committee considered the effect of a pay TV service on the existing services and on entertainment and sporting interests generally and concluded that "it is highly unlikely that a service of subscription television would significantly increase the range and quality of programming" and recommended that no service or experimental scheme of subscription television, either by wire or radio, be authorized.

While the Government "will take careful note of the Committee's arguments it recognizes that there are cogent considerations in a contrary sense. For this reason the Government for the present reserves its decision on pay-television" (White Paper).

SOUND BROADCASTING.—After considering in some detail in Chapter XVII the submissions of various bodies on the potentialities and need for "local sound broadcasting" provided from advertising revenue the Committee recommends "that one service, and one only, of local sound broadcasting be planned; that it be provided by the B.B.C. and financed from licence revenue; and that the frequencies available be so deployed as to enable it to be provided for the largest possible number of distinctive communities." The rider was added that this development of local sound broadcasting must not delay the completion of the national coverage of the three existing programmes on v.h.f. The submission by the Automobile Association for the setting up of a special radio service for motorists was not accepted. The Report does, however, recommend that the B.B.C. and the motoring organizations consult together with a view to conducting an experiment of broadcasting information to motorists on road and traffic conditions.

The Government agrees that the justification for local sound broadcasting would be the provision of a service genuinely "local" in character. The White Paper states, "As yet, however, there has been little evidence of any general public demand for this and the Government would be loath to 'create' extra demand on resources which, for the present, should be concentrated on national requirements. The Government would, therefore, prefer to take cognizance of public reaction before reaching a decision."

EDUCATIONAL BROADCASTING.—Broadly, the submissions to the Committee on educational broadcasting (which were concerned primarily with television as the medium) fell into two main groups. The first centred on the allocation of an "educational channel" with the establishment of a new national authority assuming responsibility for the service. The second envisaged an educational service within the framework of the existing broadcasting services. Having considered these proposals the Committee recommends against the introduction of a specialized service of educational broadcasting. It stresses that "the three elements, information, education and entertainment, are largely inseparable constituents of broadcasting and the production of a programme devoted to any one of them will be better for bringing with it something of the other two."

The Government concurs with these opinions and is prepared "to authorize at once additional hours for the B.B.C. and I.T.A. television services, provided these are used for programmes for the education of adults."

RELAY SERVICES.—Chapter XX of the Report considers the evidence submitted regarding the service provided by the relay companies to the million or more subscribers of whom about 50% receive television by line. It is recommended that the licences from the Postmaster-General under which the companies operate should be renewed for a period expiring on the same date (July, 1976) as the B.B.C. Charter and the Television Act. This is accepted by the Government.

At present, relay companies are obliged to ensure that their subscribers hold a wireless receiving licence. This has been a bone of contention with the relay companies for some time because no such obligation is placed upon renters of sets. The Committee recommends the exclusion of this obligation from the relay companies' licences. The Government "cannot accept" this recommendation and adds "subscribers should know that licences must be taken out and it is reasonable to expect the companies, whose function it is to supply a service which requires a licence, to see that one is held."

Other recommendations which the Government will consider "when the terms of the new licence are drawn up" include:—

The P.M.G. should continue to reserve the right, at two years' notice, to require relay companies to sell to him their assets when the licence expires.

The relay licence should require relay companies to transmit all the national sound services of the B.B.C. before relaying foreign services.

Programmes of independent television intended for one area should not be relayed in to another.

The service of continuous background music (by wire) from recordings was considered under the heading of relay services. The proposal that companies providing "music by wire" should be allotted frequencies is rejected and it is recommended that the service should be continued to be restricted to business premises.

Complementary Multivibrator

HIGH MARK-SPACE RATIOS WITH SINGLE TIMING CIRCUIT

By J. C. RUDGE,* B.Sc.

THE Eccles-Jordan bi-stable switch is one member of a closely-related family of three types of circuit: the bi-stable toggle or relay, the monostable flip-flop and the astable multivibrator. It is therefore reasonable to expect that the form of bi-stable toggle which uses a complementary pair of transistors[1] has corresponding monostable and astable configurations.

The fundamental difference between the complementary switch and the normal type (using similar-polarity transistors) is that the complementary pair conduct and are cut off simultaneously, the total current drawn varying widely throughout the cycle. The result of this situation is that only one timing circuit is required to determine both the "mark" and the "space."

Operation proceeds as follows. Assuming that T1 in Fig. 1 is at the threshold of conduction, base current flowing through R_T causes β_1 times this to flow in T2 base. The negative excursion of voltage at T2 collector is communicated via the timing circuit CR_T to T1 base, this sequence of events constituting a switching action. At the completion of the transition, T2 collector is roughly at 0V, T1 base has attempted to go to 0V, but has been held by base current in R_T to a fraction of this. Both transistors are saturated.

C now charges through the saturation resistances of the two transistors, and this charge determines the "mark." As the current flowing in T1 base decreases, the decrease in collector current initiates a reverse switching action, the results of which are that T1 base goes to +12V, and both transistors are cut off. C discharges through R_T in a time CR_T

* E.M.I. Electronics Ltd.

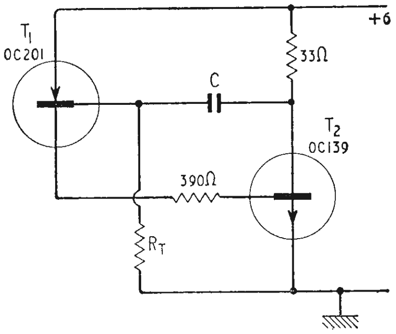


Fig. 1. Basic form of astable complementary multivibrator.

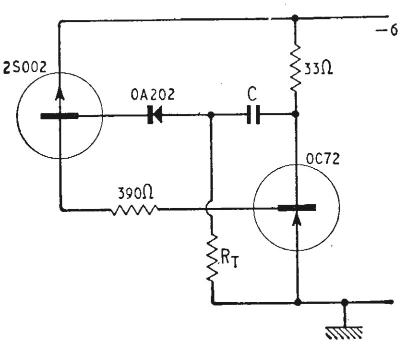


Fig. 2. Inverted form of Fig. 1. Typical values for C and R_T are 1μF and 1MΩ, and give a mark/space ratio of about 1300:1 at p.r.f. of 1.37c/s.

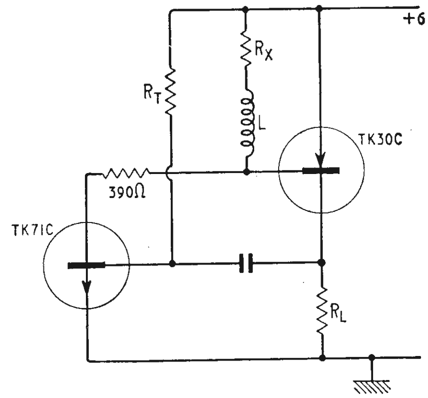


Fig. 3. Use of inductor for shortening decay-time.

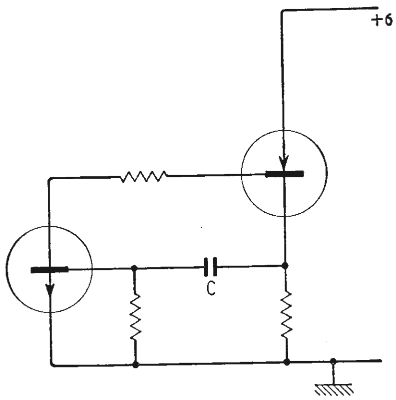


Fig. 4. Monostable circuit.

$\log_e 2$, the first transistor again conducts and the cycle repeats. The ratio of "mark" to "space" is determined by the relative values of R_T and the combined saturation resistances of the two transistors. With the transistor types shown here, ratios of 1:100 are typical and ratios of 1:1000 have been obtained.

The 390 ohm resistor between the transistors is a precautionary measure, designed to avoid a semiconductor path through the circuit, and plays no part in the operation. The diode shown in Fig. 2 is required when the circuit is used in the "inverted" configuration, the base of the silicon n-p-n transistor being fed by the timing circuit. The reverse-bias leakage of the transistor shunts the resistor R_T, and when long "space" periods are required the effect can be serious. Some silicon n-p-n transistors exhibit lower leakage and therefore need no diode.

Charge storage in the output transistor causes the familiar long decay time to the pulse, and the most successful method tried for the shortening of this long tail is the circuit shown in Fig. 3. Current in L builds up through R_X to about half the T2 base current during the pulse. The end-of-pulse step causes the back voltage in L to extract the stored charge in T2 base, thus shortening the tail.

A monostable configuration is shown in Fig. 4. The circuit is not comparable to the normal type, as the pulse length is somewhat variable, depending as it does on transistor parameters. The recovery time is long and the pulse length short.

Reference

1 *Wireless World*, May 1961, Technical Notebook, p. 280.

BOOKS RECEIVED

Ferrites—An Introduction for Microwave Engineers, by R. A. Waldron. A treatise on the microwave theory of ferrites directed at the engineer, physicist and mathematician. Some knowledge of crystal chemistry and atomic theory is required. The first two chapters on the fundamentals and general properties of ferrites are at a level which will lead the non-specialist reader gently into the more esoteric part of the book. A chapter is devoted to applications with a view to establishing a sound basis for further reading, which is suggested in the bibliography. Pp. 240. D. Van Nostrand Company Ltd., 358, Kensington High Street, London, W.14. Price 50s.

Radio Control Handbook, by Howard G. McEntee. An extremely practical book by an experienced American aero-modeller, dealing with all aspects of model control. The emphasis in this completely revised edition is on component parts, although complete systems are dealt with. Both "bang-bang" and proportional control are discussed, and it is borne in mind that the reader is a modeller first and radio engineer a bad second. While the equipment is mainly intended for model aircraft, it is easily adaptable to ships, cars and other "toys for grown-ups." Pp. 304. Gernsback Library Inc., 154, West 14th St., New York, 11. N.Y. Price $4.95.

Physics and Techniques of Electron Tubes, by R. Champeix. This first volume of a series on thermionic valve engineering is concerned with the physical principles and technology of vacuum technique. After a review of physical and chemical phenomena occurring in the gaseous state, a description is given of methods and equipment used in the establishment and measurement of low pressures, and this is followed by a discussion of miscellaneous processes employed in valve manufacture. Problems are set on each chapter. Pp. 221. Pergamon Press Ltd., Headington Hill Hall, Oxford, Price 60s.

Proceedings of the Third International Congress on Acoustics, edited by L. Cremer. A two-volume work containing reprints (in some cases shortened ones) of the 370 papers presented at the symposium in Stuttgart in September 1959. The first volume is devoted to principles, applications being contained in Vol. 2. Pp. (both volumes) 1320. D. Van Nostrand Co. Ltd., 358 Kensington High Street, London, W.14. Price 75s.

A History of Electrical Engineering, by P. Dunsheath. Written by a past President of the I.E.E., the book traces the development of electrical knowledge from the earliest, hardly-understood manifestations, to contemporary practice in power generation and distribution, and electronics. The text is enhanced by discussion of the relation of the advance of knowledge to the social scene, and by descriptions of the personalities of the men who made the discoveries. Pp. 368. Faber and Faber, 24 Russell Square, London, W.C.1. Price 50s.

Radar Pocket Book, by R. S. H. Boulding. Information on the basic principles of radar, presented in encyclopædic form, and intended for the installation and maintenance engineer or operator. Centimetric techniques are assumed throughout, and the rationalized m.k.s. system is used in the small amount of mathematics given. A chapter is included on test gear and test procedures. Pp. 248. George Newnes Ltd., Tower House, Southampton Street, W.C.2. Price 21s.

Electronics As a Career by H. F. Trewman. When a school-leaver says he wants to be an electronic engineer, he is often rather vague as to the precise meaning of the term, and this book is designed to help both parents and boys define more precisely the branch of the art that he means to follow. There is guidance on the qualifications required for all levels of employment and advice on how to set about getting them. Addresses of relevant organizations are appended. Pp. 136. B. T. Batford Ltd., 4 Fitzhardinge Street, Portman Square, London, W.1. Price 12s. 6d.

Using Transistors, by D. J. W. Sjobbema. An introduction to transistor circuit practice for the student. This second edition has been revised, and a chapter on pulse circuits has been added. The practical aspect of the subject is given rather more prominence than the mathematical side, and several circuits for amplifiers, radios, etc., are included. Pp. 128. Cleaver-Hume Press Ltd., 31, Wright's Lane, Kensington, London, W.8. Price 15s.

Flight Test Instrumentation, edited by M. A. Perry. The proceedings of the first international symposium on the subject, held at Cranfield in 1960. Fifteen papers are reproduced. Pp. 153. Pergamon Press Ltd., Headington Hill Hall, Oxford. Price 50s.

WORLD OF WIRELESS

Telstar

INTERCONTINENTAL telecommunication via an active earth satellite has been achieved. Telstar, which we described in our May issue, was successfully launched into its prescribed elliptical orbit (perigee 590 miles, apogee 3,500 miles) from Cape Canaveral, Florida, on July 10th, and experimental transmissions have been relayed from both sides of the Atlantic.

It was first used to relay television pictures between two stations in the United States. It was not until its sixth orbit that Telstar was "mutually visible," for sufficient time in the U.S.A. and in Europe for the television signal from Andover, Maine, to be picked up. At the G.P.O. station at Goonhilly, Cornwall, the first results received around midnight (10/11th) G.M.T. were encouraging but there was some frame slip and no sound accompaniment, although better results were obtained in France. On the 16th orbit, at midnight on 11/12th, a steady picture with sound was received at Goonhilly and relayed by the B.B.C. and I.T.A. During the same orbit transmissions were sent from the American designed station set up at Lannion, Brittany, these were picked up at Goonhilly and seen by British viewers as well as being received in the U.S.

We understand from the Post Office that the failure to secure satisfactory reception at the first attempts was not due to an error in tracking the satellite, but to "the reversal of a small component in the aerial feed which determined the rotation of the radio wave received." This error arose, it is stated, from "an ambiguity in the accepted definition of the sense of the rotation of the radio waves —a difficulty encountered recently in the U.S. as well as in Britain."

A series of tests covering the primary communications objectives of Telstar—telephony, telegraphy, facsimile, etc.—are now being conducted.

British Relay Pay TV

A SYSTEM of subscription television developed for possible use on the British Relay Wireless & Television Ltd. wire distribution networks was demonstrated recently in London. Distinguishing features are the simplicity of the subscribers' terminal equipment and the fact that payment is for viewing time rather than for programmes as a whole.

The meter is designed to take a single coin (2s) and accounting is carried out entirely from the central control station, which interrogates subscribers' units every three minutes and controls viewing time at rates which can be varied from 6d to 39s 6d per hour in steps of 6d. Should a subscriber wish to discontinue viewing a programme, he has only to switch off, when the unexpired value will be credited to him.

It is estimated that, because of the simplicity of the subscribers' meter it should be possible to install and maintain it for a charge of only 9d per week additional to the normal charge for the television display unit.

Colourful Radio Show.—At this year's National Radio and Television Exhibition, Earls Court, London, August 22nd to September 1st, colour television on 625 lines will be demonstrated by the industry for the first time in a "Colour Television Avenue" formed by 16 different receivers. The programme will mainly consist of colour film transmitted by land-line from the B.B.C. at Shepherd's Bush. Other technical service innovations include a black-and-white programme on 625 lines on both v.h.f. and u.h.f. The U.K. radio industry is to spend more than £30,000 on its technical services at the Show this year, state the organizers.

Educational Television Programmes Begin.—What is described as "The first University of the Air in the U.K." was introduced on July 2nd by Ulster Television, the programme contractors to the I.T.A. for Northern Ireland. Under the title "Midnight Oil," a two-month series of late evening half-hour programmes are being conducted mainly by members of the staff of Queen's University, Belfast, and cover medicine, law, literature, music, physics, history and economics. It is announced that the B.B.C. is to start, in the autumn, broadcasts in television and sound radio for technical colleges and colleges for further education.

New Structure for B.R.E.M.A.—Plans for a new structure for the British Radio Equipment Manufacturers' Association are to be put before the next a.g.m., it is announced. A. L. Sutherland, chairman, stated that all manufacturers of television receivers and nearly 100 per cent of the manufacturers of radio sets were now members of the Association.

European Television Stations now total 1,800 according to the latest (7th) edition of the "List of Television Stations" issued anually by the European Broadcasting Union. The list, together with a bi-monthly supplement, costs 50 Belgian francs a year from the E.B.U. Technical Centre, 32 Avenue Albert Lancaster, Brussels. Stations are listed both geographically and by carrier frequency. A chart showing diagrammatically the occupancy of the various channels in Bands I, III, IV, and V is also included.

A trophy to be awarded for technical efficiency, in the form of a polished bronze delta wing shape engraved with an electronic circuit pattern, has been presented to the R.A.F. No. 2 Radio School, Yatesbury, by Mullard Ltd. The trophy, by sculptor Keith Godwin, will in future be presented annually by the School to the squadron at Yatesbury obtaining the most points for technical efficiency.

Cybernetics.—The proceedings of the third International Cybernetics Conference held in Namur, Belgium, last September, are to be published in book form by the Association Internationale de Cybernétique, 13 Rue Basse-Marcelle, Namur. The 900-page volume will cost 1,000 Belgian francs if ordered before August 31st, after which it will be 1,300 francs.

Colour TV Receiver Servicing.—A course of six lecture-demonstrations covering both the NTSC and SECAM systems are to be given by B. J. Rogers, of Bush Radio, at the Norwood Technical College, London, S.E.27, on successive Tuesday evenings from October 9th. The fee is 15s.

Over-the-air subscription television had its premier in North America on June 29th when R.K.O. General station, WHCT, Hartford, Connecticut, broadcast the motion picture "Sunrise at Campobello" on u.h.f. channel 18 to about 300 Hartford families. Phonevision equipment, developed by Zenith Radio Corporation, is used and this is progressively being installed in further subscribers' homes. The station is licenced to conduct a three-year test.

A Band IV television aerial is to be mounted above the existing Band I aerial at the B.B.C.'s Crystal Palace television station. The aerial will be omni-directional, horizontally polarized and of high gain. It will consist of eighty elements of end-fire stacked dipoles mounted in angled fashion from the corners of the tower, and will have a bandwidth which will cover several television channels. It is planned to have the aerial, which is being supplied by Marconi's W/T Company, available for use early in 1963.

Ghosting and attenuation, roughly along a line from Croydon to Southend, is resulting from construction of a new mast at the Independent Television Authority's Croydon station. This interference to the radiation from the existing aerial is inevitable, state the I.T.A., now that the mast has reached a critical height. The new mast and aerial will be brought into service in the autumn.

New Chairman for B.V.A.—F. V. Green, who has represented Brimar interests on the board of management of the British Radio Valve Manufacturers' Association since 1947, has been elected chairman as from July 1st. He succeeds J. W. Ridgeway, O.B.E., who has retired from industry. Mr. Ridgeway was chairman of B.V.A. for 18 years, holding office since July 1942, with the exception of a two year period (1947-49) when George Marriott was in the chair.

Royal Society Grants.—An award of £2,736 for the development of an instrument for measuring the depth of continental ice sheets by a radar technique, has been made to Dr. S. Evans, senior assistant in polar research, University of Cambridge, by the Paul Instrument Fund Committee. As a result of an award under the Royal Society and Nuffield Foundation Commonwealth Bursaries Scheme, Professor E. L. Yates, professor of physics, University College of Rhodesia and Nyasaland, is at present on a five-month visit to Melbourne discussing problems of solid state physics.

Computer Film.—Among new sponsored films which are available to users without hire charge from the Rank Film Library, 1 Aintree Road, Perivale, Greenford, Middx., is one by I.C.T. Ltd. entitled simply, "The 1301." The film shows the work capacity and applications of this electronic computer, and is intended specifically for the information of management engaged in the scientific control of affairs and for professional bodies concerned with the development of data processing. The colour film, reference number 91/6410, lasts 28 minutes.

Two new training courses for potential and practising works managers in place of the existing training course, will be inaugurated in September by the Institution of Works Managers, 196 Shaftesbury Avenue, London, W.C.2. The Certificate Course will be open to candidates anywhere, whilst the higher standard Diploma Course will be held at approved colleges.

Wesley Evening Institute, London, N.W.10, will again be holding a Radio and Television Course, intended primarily for amateurs, during the 1962/63 session. Details are obtainable from the head of the Institute—E. N. Fennell, "Jeanville," Brighton Road, Addlestone, Surrey.

12M Licences?—An increase in May of 62,558 combined television and cound radio licences issued throughout the U.K. brought the total up to 11,928,909. G.P.O. figures for June and July are not yet available, but it is probable that the total now exceeds the 12M mark. Sound only licences totalled 3,508,208, including 508,050 for sets permanently fitted in cars. In comparison there were 6,493,411 registered television sets and 16,481,314 sound radio receivers in West Germany and in W. Berlin on May 1st. The figure for television represents an increase of 101,458 over the previous month's total.

Radio Heliograph.—The Commonwealth Scientific and Industrial Research Organization of Australia is to build a radio heliograph consisting of 100 paraboloids, each 42ft across, linked to receivers, the outputs from which will be fed into computers. The aerials will be situated around the periphery of a two-mile diameter circle. Designed to give a "moving picture" of the radio flares which accompany sunspot explosions, the building of the heliograph has been made possible by a gift of 550,000 dollars from the Ford Foundation. It will probably be built at Parkes, N.S.W., where the C.S.I.R.O.'s 210ft radio telescope is sited.

Italian Radio Show.—The 28th annual Italian radio and television exhibition (28a Mostra Nazionale Radio-Televisione) is to be held at the Palazzo dello Sport, Milan, N. Italy, from September 5th to 12th inclusive. The programme for this year's show includes a 3-day technical meeting on electronic components commencing on September 10th. English speaking visitors are being especially welcomed on September 9th with a reception arranged by the exhibition organizing committee, followed by a detailed tour of the exhibition with interpreters in attendance.

900-mile Aerial.—Plans are being made to run an aerial along the 900-mile length of three toll roads in the American mid-west to provide a radio news service for motorists. It is proposed to use a frequency at the lower end of the medium-wave band. Radiation, it is said, will be limited to a radius of 150 feet from the aerial.

German radio and television set manufacturers are again to participate in the Deutsche Industrie-Ausstellung (German Industrial Fair), which will be held this year in West Berlin from September 22nd to October 10th.

Compatibility Test?—"On colour TV there will be spectaculars like The Black and White Minstrel Show. . . ."—*Daily Mail*, July 6th.

SEPTEMBER ISSUE

Show Guide

Next month's issue, which will be published a week earlier than usual—on 21st August (Preview Day)—will contain a stand-by-stand guide to the Earls Court Radio Show together with a plan and list of exhibitors.

This will be in addition to the normal quota of pages devoted to regular features and technical articles.

OCTOBER ISSUE

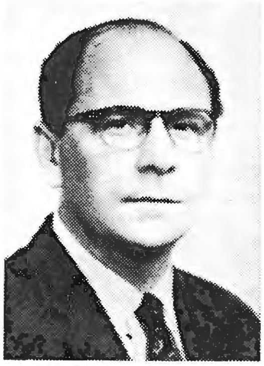
Show Reviews

In this issue the Technical Staff of *Wireless World* will give their impressions of the year's trends in sound and vision broadcast receivers as exemplified at Earls Court, and aeronautical electronics as seen at the Farnborough Air Show. It is hoped to include reports on the Fourth International Congress on Acoustics at Copenhagen and the Microwave Valves Congress at The Hague.

Personalities

P. E. Trier, M.A., M.I.E.E., M.Brit.I.R.E., director of Mullard Research Laboratories, has been appointed to the board of Mullard Ltd. as from June 1st. Mr. Trier, aged 42, was educated at Mill Hill School and graduated as a wrangler in the Mathematical Tripos at Cambridge. He was engaged at the Admiralty Signal and Radar Establishment from 1941 to 1950, and did research in u.h.f. and microwave techniques. During the latter part of his service with A.S.R.E. he was head of the V.H.F. Communications Group. Mr. Trier joined Mullard Research Laboratories in 1950 as head of the Communications and Radar Division. He was appointed manager of the Laboratories in 1953 and director in 1957.

P. E. Trier



J. D. Clare

T. R. Scott, D.F.C., B.Sc., M.I.E.E., has relinquished his position as managing director of Standard Telecommunication Laboratories Ltd., an S.T.C. subsidiary. Mr. Scott will continue his current activities in the power cable field where, *inter alia,* he represents S.T.C. on the Council of the Cablemakers' Association, of which body he has recently been elected vice-chairman. He will also continue as a member of the board of directors of Enfield-Standard Power Cables Ltd. Mr. Scott is succeeded at Standard Telecommunication Laboratories by **John D. Clare,** M.Sc., A.M.I.E.E., who, after an initial period in industry with G.E.C. and Sobell, joined the Radar Research Establishment and eventually was promoted to Ministry of Aviation headquarters where he was, until his present appointment, director of research, guided missiles.

Robin W. Addie, M.A., marketing director of E.M.I. Electronics Ltd. since 1958, has resigned from the company to join the board of Painton & Co. Ltd. Mr. Addie was for some years technical commercial manager of Philips before joining E.M.I. Electronics in 1957 as export manager.

Commander F. Holmes has been appointed managing director of Westrex Co. Ltd. This is in succession to **H. L. Marsterson** who resigned, due to ill health, earlier this year. Commander Holmes was appointed sales director of Westrex in August, 1961, and since November has been acting managing director.

W. G. Patterson, M.B.E., divisional director and general manager of A.E.I., Telecommunications Division, has been elected chairman of the Telecommunication Engineering and Manufacturing Association. **C. Riley,** director and general manager of the General Electric Co. (Telecommunications), becomes vice-chairman of the Association.

Denis Taylor, Ph.D., M.Sc., M.I.E.E., who for the past five years has been a director and general manager of Plessey Nucleonics Ltd., has been appointed co-ordinator of research and development for the Plessey Co. Ltd. At one time head of the Electronics and Instrument Division of the Atomic Energy Research Establishment, Harwell, Dr. Taylor, who is 52, has held a number of lectureships in electrical engineering and physics. In 1939 he joined the Air Ministry as a scientific officer and later played a leading part in the development of radar. A former member of the Council of the I.E.E. and past chairman of the Measurement and Control Section, Dr. Taylor is the chairman designate of the New Science and General Division of the Institution.

G. D. Cook, A.M.I.E.E., has been appointed by the B.B.C. as regional engineer, Wales, in succession to **J. E. F. Voss,** B.Sc., A.M.I.E.E., who has recently been appointed superintendent engineer, television (London studios). Mr. Cook joined the Corporation in 1947, as a maintenance engineer at the Brookmans Park transmitting station. He transferred to the Transmitter Section of the Planning and Installation Department in 1949 and to the Television Service in 1955 when he became assistant to the superintendent engineer, television (regions and outside broadcasts). In 1959 Mr. Cook was appointed engineer-in-charge (television), Manchester.

The B.B.C. announces the retirement of **J. H. Holmes,** B.Sc., M.I.E.E., A.C.G.I., superintendent engineer, lines, and the appointment of **G. Stannard,** B.Sc., M.I.E.E., A.C.G.I., to succeed him. Mr. Holmes joined the B.B.C. in 1935 as an assistant engineer in the lines department. He was successively promoted in the department and became superintendent engineer, lines, in 1947. Mr. Stannard began his B.B.C. career in 1932 as a maintenance engineer in the London Control Room and in 1933 he transferred to the section working on sound recording. He moved to the lines department in 1935.

The B.B.C. has appointed **W. D. Hatcher,** B.Sc.(Eng.) to the post of assistant superintendent engineer, television (London studios), to succeed **M. H. Hall,** who has retired. Mr. Hatcher joined the B.B.C. in 1931 and since 1960 has been engineer-in-charge, television studios.

C. B. Flindt, B.A., A.M.I.E.E., who joined the development division of R. B. Pullin four years ago, has recently been appointed chief engineer of the company under **Dr. R. H. Barker,** the newly appointed technical director. Dr. Barker's appointment followed the resignation of **Dr. Sydney Jones,** who left R. B. Pullin to become Director of Research in British Railways.

R. R. Kennedy, B.Sc. (Eng.), A.M.I.E.E., chief engineer of S.E. Laboratories (Engineering) Ltd. for the past four years, has been appointed technical director of the company. Before joining S.E. Laboratories Mr. Kennedy was a systems engineer at English Electric working on ground-to-air guided missiles.

Dawson Donaldson, at present Director-General of the New Zealand Post Office, is to succeed **Sir Ben Barnett,** K.B.E., C.B., M.C., as chairman of the Commonwealth Telecommunications Board. Mr. Donaldson, who is 59, is likely to take up his appointment in November on Sir Ben Barnett's retirement.

Two senior Marconi engineers, **Geoffrey E. Beck,** B.Sc. (Hons.), and **Mervyn Morgan,** M.Sc., have been awarded the Johnston Memorial Trophy by the Guild of Air Pilots and Air Navigators for their work on the development of the airborne Doppler Navigator for both military and civil aircraft. Mr. Beck has been working on the development of Doppler Navigators since 1951 and now holds the position of chief engineer (navigational aids) in Marconi's Aeronautical Division. Mr. Morgan has been associated with the development of Doppler navigation since 1950 and is at present superintendent, Communications and Navigational Aids Research Laboratories, at the Marconi research establishment near Chelmsford.

Martin T. Mason has joined 20th Century Electronics as sales manager. After a period with E.M.I. he was, for many years, a senior engineer at Dollis Hill Research Station where he was concerned with u.h.f. communications, piezoelectric crystals and with early developments of radar. He also established laboratories for work on electronic tubes and high vacuum techniques. After the war Mr. Mason joined Edwards High Vacuum and from 1950 to 1952 was chief designer for K.D.G. Instrument Ltd. He spent eight years with Salford Electrical Instruments (G.E.C.) on the design and sales of industrial instrumentation and for the past two years has been technical sales manager of Balzers High Vacuum Ltd.

J. H. Barker, formerly sales director, has been appointed marketing director of Bush Radio, a division of the Rank Organisation. A number of executive changes have followed. **Colin Taylor,** who has been with the company for 27 years, is now home sales manager, and **John Hignett** is the new publicity manager. Mr. Hignett joined Bush Radio some months ago from Texas Instruments and before that he was with Pye Telecommunications. **G. P. Wickham Legg,** who has been publicity director since 1952 and has served Bush Radio since the formation of the company in 1932, has retired, but is remaining with Bush in an advisory and consultative capacity until the end of the year.

Ian Band, B.Sc., a member of the Hughes International (U.K.) Ltd. research engineering team at Glenrothes, Fife, since 1960, has gone to Newport Beach, California, headquarters of the semiconductor division of Hughes Aircraft Company, to act as liaison engineer between the headquarters and the U.K. organization. Mr. Band, a graduate of Queen's College, St. Andrews University, was a radar research engineer with Ferranti, Edinburgh, before joining Hughes.

The B.B.C. announces the appointment of **D. P. Leggatt,** B.Sc., as engineer-in-charge, television recording, to succeed **R. S. Meakin,** A.M.I.E.E., who has been appointed superintendent engineer, television (recording). Mr. Leggatt joined the B.B.C. in 1953, serving for two years in the Engineering Information Department, followed by four years in the Planning and Installation Department. In 1959 he transferred to the Television Operations and Maintenance Department.

Frank Hicks-Arnold has recently been appointed general manager of Bribond Printed Circuits of Chichester. After being closely concerned in the introduction of printed circuits to this country with John Sargrove Ltd., he spent several years as a development engineer with A. C. Cossor Ltd. A lecturer on printed circuit techniques, he read a paper on automation to the Cambridge University Convention in 1958 and last year was a U.K. delegate to the International Electrotechnical Commission.

Air Vice-Marshal W. L. Freebody, C.B., C.B.E., A.F.C., i.d.c., has been appointed managing director of RCA Great Britain Ltd.

H. W. L. Cumming, B.Sc., has recently been appointed assistant manager (semiconductors) of the Valve and Semiconductor Engineering Department, A.E.I. Electronic Apparatus Division, Lincoln. He was a laboratory apprentice at Siemens Electric Lamps & Supplies Ltd. from 1940 to 1943. After holding various laboratory appointments he was appointed section leader of the Transistor Section at Siemens Ediswan Research Laboratory in 1956 and since 1957 has been chief engineer (semiconductors), of A.E.I. (Woolwich) Ltd.

T. A. Connor, B.Sc.(Eng.), has been appointed chief engineer in charge of research, development and design at the new Bognor Regis Components Division of Royal Worcester Industrial Ceramics Ltd., of Tonyrefail, South Wales, the makers of the Regalox ceramic for industrial use. He served two years graduate apprenticeship with Sperry Gyroscope Co. Ltd., continuing as an engineer with them on guided missile development from 1952 to 1957 when he left to join K.L.G. Sparking Plugs Ltd., as the engineer responsible for the design and development of sealed terminals. **I. C. Walker,** Assoc.Brit.I.R.E., has been appointed sales manager of the Components Division of Royal Worcester Industrial Ceramics. During the war Mr. Walker was employed in the laboratories of Rediffusion Ltd., on synthetic d.f. and radar trainers. In 1948 he went to Rediffusion (Malta) Ltd. He joined S. Smith & Sons Ltd. in 1950 in the Radiomobile car radio division and was engaged in engineering liaison work until transferring to K.L.G. in 1955.

E. J. Blythe, B.Sc., A.M.I.E.E., has been appointed chief product engineer of Aircraft-Marine Products (Great Britain) Ltd. After a graduate apprenticeship with English Electric, he joined Ferranti and subsequently went to Standard Telephones & Cables and then to the M.O. Valve Company on the development and production of transmitting valves. The company also announce the appointment of **A. Fowke,** D.F.H., Grad.I.E.E., in charge of the A-MP laboratories. He was at one time with Metropolitan Vickers.

A. C. Cossor Ltd. announce the appointment of **Leonard H. Weall** as director of manufacturing for their group electronic companies. Mr. Weall worked his way from an apprentice to development and liaison engineer, production engineer, chief inspector and on to production manager of the Plessey Company's Rotherham Works. Before joining Cossor's Mr. Weall was works manager of G. & E. Bradley Ltd.

OBITUARY

Sir Allen George Clark, chairman and managing director of the Plessey Company, died on June 30th, aged 63. Sir Allen entered the Plessey Company in 1920, became joint managing director in 1925, and was appointed chairman in 1945. He was, at one time, a radio industry representative on the Radio Rearmament Advisory Committee, which was set up to facilitate liaison between the Ministry of Supply and the radio industry on matters affecting defence. He was awarded a knighthood in the Queen's Birthday Honours List for 1961.

T. Walmsley, C.B.E., Ph.D., B.Sc., M.I.E.E., M.I.R.E, died on June 20th at the age of 76. Dr. Walmsley joined the G.P.O. Engineering Department in 1936 as a staff engineer. In 1940 he became deputy director of communications development, Air Ministry. He was awarded the C.B.E. in 1946 and in the same year was appointed director of research, British Telecommunications Research Co., which had been formed by B.I.C.C. and A.T.E., with the object of concentrating research in telecommunications equipment. He retired from that position in 1949 and was latterly a consultant to Automatic Telephone & Electric Company.

News from Industry

E.M.I. Electronics Reorganization.—A major reorganization of the management structure of E.M.I. Electronics has taken place and all the company's commercial products have been grouped into product divisions. Each division is controlled by a product manager responsible for design, production and marketing of all products within that division. Consequent upon the reorganization the following appointments have been made: F. G. Helps, manager, industrial division; D. G. Ashton Davies, manager, instrument division; J. Sharpe, manager, valve division (commercial); C. Coles, manager, administration and finance division; C. Dain, deputy to A. H. Cooper, technical director; R. W. Divall, deputy to F. H. Panter, works director; W. S. Challis, manager, Feltham, to succeed P. W. S. Valentine who is retiring shortly; D. Rowson, deputy manager, Feltham; F. R. Trott, chief engineer, commercial; and A. J. Perera, management accountant.

Intertherm Ltd. has been formed jointly by Redifon, the principal manufacturing company in the Rediffusion Group, and Research & Control Instruments, a subsidiary of Philips Electrical Industries Ltd., to pool the manufacturing and marketing activities of the two companies in the fields of induction and dielectric heating equipment. The new company commenced operations on July 1st at Brixton Works, Blenheim Gardens, Brixton Hill, S.W.2 (Tel.: Tulse Hill 9531) and has assumed responsibility for guarantees and service on all Philips and Redifon electro-heating equipment previously supplied.

Plessey group reorganization has resulted in a new subsidiary, Plessey Co. (U.K.) Ltd., being formed. The Plessey Co. Ltd. becomes a holding company and will have seven direct subsidiaries. These are, Plessey Group Management Ltd., the Plessey Co. (U.K.) Ltd., Automatic Telephone & Electric Co. Ltd., Ericsson Telephones Ltd., Garrard Engineering & Manufacturing Co. Ltd., Semiconductors Ltd. and Plessey International Ltd.

Morganite Resistors Ltd. announce that arrangements have been made with the Board of Trade for the building of a 50,000 sq ft factory on the Bede Trading Estate, Jarrow. The company manufacture components for the radio, television, electronic and light electrical industries. This new factory—their third—will take over the manufacture of the special types of resistors now made at the Morgan Crucible Group's factory at Norton, Worcester, and will thus concentrate at Jarrow their complete range of electrical resistance products.

Pakistan is to have further installations of Marconi high-frequency radio communications equipment as CENTO aid. The equipments, which include transmitters, receivers and radiotelephone terminals, are to be installed at the Pakistan Posts & Telegraphs Department's stations at Pipri, Ghaggar, Karachi, Rawalpindi and Dacca. Marconi's are also supplying a siderable quantity of CENTO aid radio communications equipment to Turkey and Iran.

Bridlington Relay Ltd., which has operated the radio relay service in the town for 28 years, is installing E.M.I. Electronics' wide-band equipment to provide a wired television system.

Ekco Electronics Ltd. have received from the Ministry of Aviation a substantial contract to develop a special radar equipment for helicopters.

Electron-Physical Instruments Ltd. is to join the Hilger & Watts group of companies. E.P.I. was formed in 1960 by R. V. Ely, M.I.E.E., M.I.Mech.E., formerly managing director of Foster Transformers and a group director of the Lancashire Dynamo Group, who will continue as vice-chairman. The company specializes in point-focus electron beam instruments. Under the terms of the arrangement E.P.I. will continue to operate as a small company in the group carrying out the development of instruments, the manufacture of prototypes and equipment to individual requirements. Sales of these instruments will be taken over by Hilger & Watts, at 98 St. Pancras Way, London, N.W.1.

Radio & Television Trust Ltd.—Consolidated group profit for the year to March 31st, 1962, amounted to £433,466 as compared with £265,782 for the previous year. The charge for taxation is £215,265 (£127,266 for the preceding year) leaving a balance of profit for the year of £218,201. The directors of Radio & Television Trust announce that they have recommended the name of the company be changed to Controls & Communications Ltd.

Rediffusion's group trading profit and investment income for the year to March 31st, 1962, expanded from £7.1M to £7.6M. The net profit is up from £1.8M to £2M.

Fisher's U.K. Agent.—The wide range of audio equipment made by the Fisher Radio Corporation of Long Island City, New York, is now available from Lockwood & Co. (Woodworkers) Ltd., of Lowlands Road, Harrow, Middx. (Tel.: Byron 3704), who have been appointed agents and are able to deal with enquiries for all types of Fisher equipment.

Walmore Electronics Ltd., of 11-15 Betterton Street, Drury Lane, London, W.C.2, have been appointed the exclusive U.K. distributors for McCoy Electronics Co., Pennsylvania, U.S.A. McCoy specialize in quartz crystals and crystal filters. Of particular interest are their crystals for frequencies up to 200 Mc/s and their micromodule crystal units.

Hellermann Equipment Ltd., a new subsidiary company of Hellermann Ltd., of Crawley, Sussex, has been formed to take over the marketing operations of the company's Industrial Products Division.

Irish Army Air Corps has ordered a Decca 424 airfield control radar for installation at its Baldonnel military air base near Dublin.

CHANGES OF ADDRESS

The Micro Switch Division of Burgess Products Co. Ltd., is now located at Eastbury House, Albert Embankment, London, S.E.1 (Tel.: Reliance 7871). The London office of Burgess's Acoustical Division is now 127 Victoria Street, London, S.W.1.

Gas Purification & Chemical Co. Ltd. have moved their London offices to North-West House, 119-127 Marylebone Road, London, N.W.1. (Tel.: Ambassador 6671). Associated with this move will be three principal companies of the group—A.B. Metal Products Ltd., Smart & Brown (Connectors) Ltd. and Wolsey Electronics Ltd.

R. H. Cole (Overseas) Ltd. have moved their offices and showrooms to a new building at 26-32 Caxton Street, London, S.W.1. The telephone number remains unchanged as Sullivan 7060.

Electroluminescent Devices

LAMPS, RELAYS, LIGHT AMPLIFIERS AND DIGITAL INDICATORS

By L. H. BRACE*

THE phenomenon of electroluminescence in phosphors has been known for many years. Electroluminescence can be defined as the light given off by a suitably treated phosphor as a result of the interaction of a changing electric field with the phosphor. This definition describes what is known as "intrinsic" electroluminescence.

There are other ways in which phosphors may be made to electroluminesce. One of these is the effect of an electric field on crystals in contact with conducting electrodes so that current injection can occur (carrier-injection electroluminescence). The other involves an electric field and a simultaneous photo-excitation of a phosphor with the resultant emission of light (electrophotoluminescence).

Electroluminescence as we know it to-day was first discovered by G. Destriau in 1936. The phenomenon remained a curiosity solely of laboratory interest until 1949 when the first practical lamp using electroluminescence was produced.

One way in which the effect may be demonstrated is with a suitable phosphor powder mixed in a binder, such as polystyrene and spread between two transparent conducting plates. These plates may be a glass having a transparent layer of tin oxide on one surface (Nesa glass). This arrangement forms a parallel-plate luminous capacitor with the phosphor binder acting as the dielectric.

When a changing electric field is applied to this "luminous capacitor," light is emitted. The brightness and colour of this light depend upon the amplitude and frequency of the applied voltage, the temperature, composition and thickness of the phosphor material.

The maximum brightness that can be obtained is limited by the breakdown voltage of the dielectric. A thicker dielectric material, or one having higher breakdown voltage characteristics, makes it possible to increase the voltages to higher values and obtain more light. However, there is an obvious limit to the thickness of the dielectric beyond which it begins to absorb an appreciable amount of light. The best solution is to use a thin dielectric with a high breakdown voltage and dielectric constant.

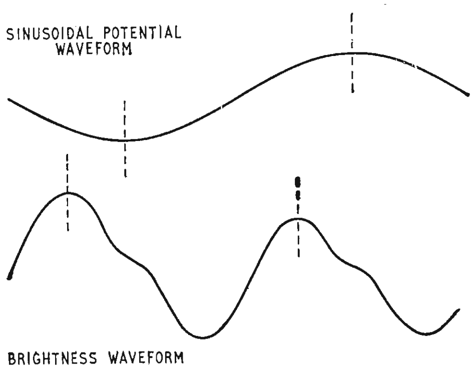


Fig. 1. Brightness waveform and sinusoid exciting voltage compared: peaks are indicated by dotted lines. Note secondary peaks "halfway down" brightness wave.

The intensity of electroluminescence also varies with the frequency of the applied voltage. Light pulses from an electroluminescent cell occur at twice the frequency of the applied voltage. When a phototube is placed near an electroluminescent cell that has a sine-wave voltage applied, the resultant output of the photocell displayed on an oscilloscope is known as a "brightness wave."

If the voltage applied to the electroluminescent cell is also displayed with the brightness wave, a peak in the brightness wave is seen to occur slightly before the positive and negative peaks of the sine wave (Fig. 1). One of these peaks may be higher than the other. At low frequencies, the brightness varies between some peak value and zero; but, as the frequency increases, the light output between the two peaks of brightness during an a.c. cycle does not fall to zero. Instead, the higher the frequency, the higher is the continuous component between peaks.

From this, it appears that light is generated for every charge and discharge of the electroluminescent capacitor; therefore, the more often the field can be reversed, then the larger will be the quantity of light emitted. There is a limit to this gain: as the frequency is increased further, a point is reached where the brightness curve levels off to a saturation value.

Not only is the brightness changed by a variation in the frequency of the applied voltage, but so also is the colour of the emitted light. For instance, light from a given cell may appear green at 60c/s, but this colour will change to a pale blue when the frequency of the applied voltage is increased to 1,000c/s.

Many theories have been proposed for explaining this phenomenon of electroluminescence; one is that electroluminescence begins with a freeing of electrons from donors under the action of the electric field. This is followed by an acceleration of the electrons in which an avalanche may multiply the original number of electrons. Collision of these electrons with "activator centres," which are thereby excited or ionized, is responsible for the production of light.

Solid-state Amplifiers

A solid-state light amplifier can be constructed, using electroluminescent and photoconductive materials in a sandwich-like construction as illus-

*Static Devices Co., El Paso, Texas, U.S.A.

trated in Fig. 2. A thin layer of photoconductor is placed on Nesa glass: on top of this is spread an opaque layer of conducting lamp-black suspended in an epoxy resin; this layer is followed by a thin film of the electroluminescent material suspended in a plastics binder. Finally, a second plate of Nesa glass is placed in contact with the electroluminescent film. The opaque layer of lamp-black prevents light feedback from the electroluminescent film to the photoconductor.

This association of photoconductive and electroluminescent materials forms an electrical circuit when a voltage source is applied. Such a circuit can be reduced to an equivalent variable resistance and fixed capacitance as shown in Fig. 3.

When the photoconductor is not illuminated, it has a high resistance (about 0.1 to 1MΩ). Consequently, if an alternating voltage is applied across the series RC circuit, most of the voltage will appear across the photoconductor.

When the photoconductor is illuminated, its resistance drops sharply to between 100 and 1,000Ω, causing the applied voltage to shift from this low resistance to the higher impedance presented by the capacitance of the electroluminescent layer. This increase of alternating voltage across the electroluminescent portion of the device provides the field which is necessary to activate the phosphors, which then emit light.

Early light amplifiers of this type were made with a thick photoconductor layer, which was necessary to control adequately the electroluminescent layer. However, because light could not penetrate all of this layer, the bottom portion of the layer acted as a resistance which severely limited the light gain and resolution.

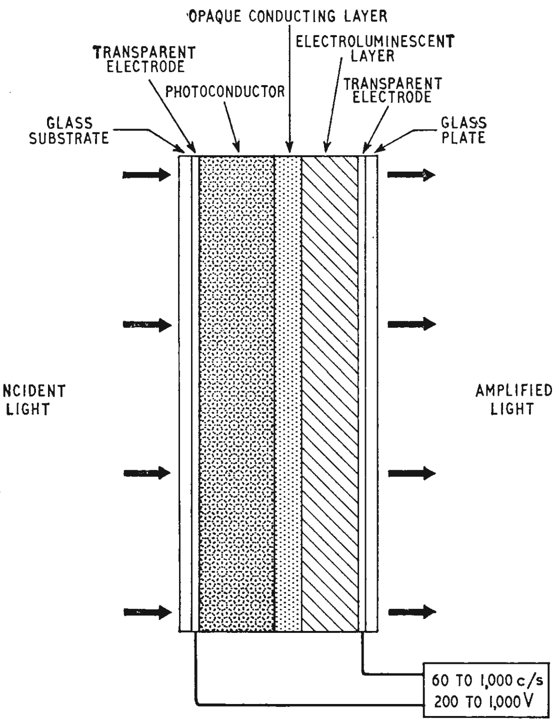


Fig. 2. Cross section of solid-state light amplifier.

This problem was solved by a fabricating process which involved the cutting of fine "V" grooves into the photoconductor. When the photoconductor is illuminated, the photocurrent that develops moves down the sides of these grooves and disperses into a phosphor layer which spreads the current in the area under the ridge of photoconductor. This layer permits almost all of the phosphor to be excited instead of only the area immediately under the groove.

The gain of this grooved photoconductive amplifier can be further improved by the arrangement shown in Fig. 4. The photoconductive layer operates more efficienly using d.c. or rectified a.c. whereas the electroluminescent layer operates only when alternating voltage is used. By connecting alternate strips of the photoconductor to one bias supply, and the rest of the strips to another bias supply of the same value but opposite polarity, the biased and unbiased alternating voltages are available to both layers simultaneously. With such an arrangement, more than 50 times as much light is obtained compared with the conventional a.c. operation.

As the amount of light emitted from the electroluminescent phosphor is directly related to the magnitude of the field, this value will vary with the intensity of the image falling on the device. Such an image will be of one colour, which depends upon the phosphor used, and the image intensity from point to point will show half-tone as well as "black-and-white" variations. The use of this R-C combination as a light amplifier depends on the control characteristics of the photoconductive material.

The photoconductor acts in much the same manner as the control grid of an electron tube; the amplifying characteristics of the device are a result

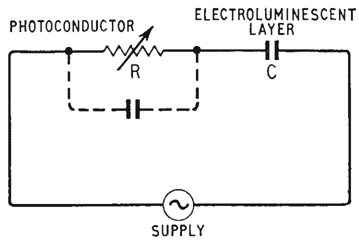


Fig. 3. Equivalent circuit for solid-state light amplifier.

of the resistance illumination characteristic of the photoconductor with the additional energy in the emitted light supplied by the electric field between the Nesa-glass plates.

Solid-state Light-activated Relays

The two major components of the solid-state light amplifier—the photoconductor layer and the electroluminescent layer—can be arranged for use in logic systems as relays, scale-of-two counters, and shift registers. Of these, the relay is the most easily constructed and simple to operate.

Fig. 5 shows the arrangement of elements in the solid-state relay, and compares its mode of operation with a conventional electro-mechanical relay.

When light is emitted from the electroluminescent layer of the solid-state relay, the resistance of the photoconductor will drop, allowing current to flow in the circuit in which it is a controlling element.

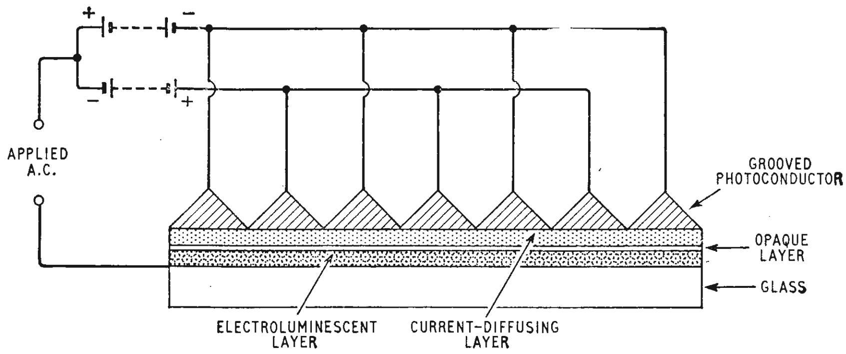


Fig. 4. A.c. operation of light amplifier of grooved type.

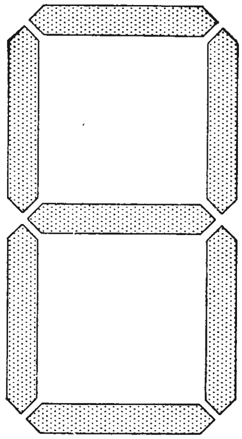
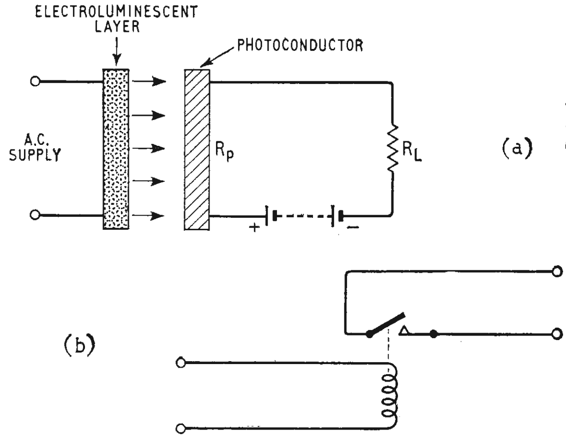


Fig. 6. Digital indicator segment pattern. Any number between zero and 9 can be formed by use of part of pattern.



Left: Fig. 5. (a) Representation of solid-state relay compared with (b) mechanical relay.

This is equivalent to closing the contacts in a conventional electro-mechanical relay. The advantage of the solid-state relay is that there are no moving parts to stick or contacts to spark as is the case with conventional relays.

When the conventional electro-mechanical relay is open, the insulation resistance between the contacts is very high (hundreds of MΩ); with a resistance of only a few milli-ohms when closed. In comparison, the open condition of the solid-state relay is about 300Ω/cm² of active area, and about 3,000Ω/cm² when closed.

It is possible to use a number of emitters to simulate a multicoil relay; conversely, a single emitter can operate a number of photoconductor elements arranged suitably about the emitter. Good separation of control and load terminals is provided because there is no need for physical contact between electroluminescent emitter and photoconductor. Also, it is possible to determine visually the condition of the relay—when the emitter is illuminated, the relay is closed.

Radar and Television Applications

The principal element in both television and radar for information display is the cathode-ray tube. The electron beam must not only carry the information to be displayed but must be of sufficient intensity to excite the screen phosphors to a level that will make the information clearly visible.

The room in which the radar or television screen is sited usually has to be darkened below normal room illumination. By replacement of the phosphor screen with a solid-state light amplifier, or the use of such an amplifier in contact with the screen, a reduction in the electron-accelerating voltage and control circuitry required can be realized. Information presented to the solid-state light amplifier is then amplified for display by the relatively low voltages required to operate this device.

Using this latter mode of operation, it is possible to place in contact with the phosphor screen a grooved-type light amplifier which, if the output of the screen is matched with the point of peak response of the photoconductor, will show an integrated energy gain of about 200. For an input of one lumen, an output of between 100 and 200 lumens is possible. As the ambient illumination in laboratories and homes lies between 50 and 80 lumens/ft², the output of this light panel is sufficient to see the image clearly.

Light amplifiers of this type have been developed in sizes greater than one foot square, in thickness only slightly greater than the supporting glass plate and the resolution achieved is over 500 television lines.

Digital Number Display

The displaying of information directly as figures or letters is becoming popular because it gives a positive indication which cannot be misread as can a pointer-type (analogue) instrument. The digital instrument is compact and increases the ease with which information can be read—there are no adjacent scales or numbers to confuse or impede the speed with which an observer can recognize and interpret information.

The electroluminescent display consists of a single panel about a quarter-of-an-inch thick, which contains the pattern of the digits to be displayed. Panels can be made with digits ranging from one foot in size down to eighths of an inch; also, each panel can contain a large number of digits limited only by the complexity of the circuit and switching arrangements.

With the 7-segment digit shown in Fig. 6, any number from 0 to 9 can be formed by connecting the appropriate combinations of segments. This is easily seen by covering the segments unnecessary to the formation of any given numeral.

There is no parallax problem, the panel can be easily viewed at wide angles, the numerals are formed in a single plane so that the segments do not mask each other. By varying the frequency of the excitation signal, it is possible to vary the colour of the digits; in cases where space is at a premium, the panel itself can be mounted where convenient and the switching equipment placed where there is more room.

"FILTERNICS"

By THOMAS RODDAM

ACTIVE FILTERS WITH NEGATIVE IMPEDANCE CONVERTERS

LOOKING back in the files, and if anyone wants to know what my files look like the answer is "hellish dark, and smelling of cheese", I see that at various times I have discussed negative impedance amplifiers and the use of feedback amplifiers as filters. The invention of the negative impedance amplifier has led to a whole new field of network design with the inductance and capacitance systems replaced by capacitance, resistance and transistrance. This last monstrous word, the property of being a transistor, is copyright and the full rigours of the law will be invoked against anyone using it. So far all the discussion of these systems has been on a rather rarefied plane and the design theory is likely to remain there, but the fact that we cannot carry out the detailed computation is no reason why we should not know what these systems look like. This article is a general survey of the sort of circuits which are being used.

We must go back to the beginning of the story. It was shown by Merrill (*Bell System Technical Journal*, Jan. 1951, p.88) that a particular valve circuit could be designed to give, between two terminals, an impedance which satisfied the equation

$$V = -IR$$

for signal voltages and currents over a reasonably wide frequency range. More important, R could be made to be something like $[\mu/(\mu + 1)]\, R_1$, where R_1 is the resistance of a physical resistor and μ is a valve amplification factor. In consequence, $-R$ is rather well defined and can be used as a circuit element. One use is to connect $-R$ in the path between a generator of impedance R_0 and an equal load. The load current becomes $V_0/(2R_0 - R)$ instead of $V_0/2R_0$ and we have an apparent gain

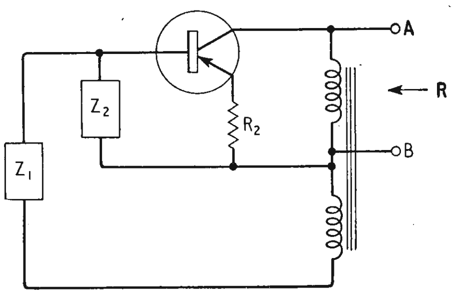


Fig. 1. Basic transistor inverter circuit.

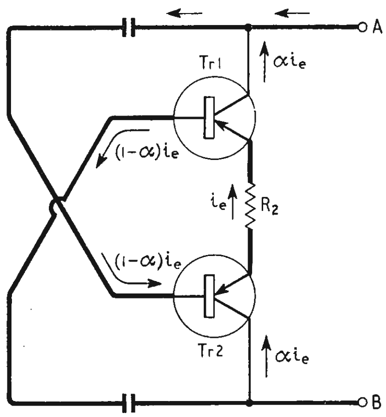


Fig. 2. The use of two transistors permits the elimination of the transformer in Fig. 1.

of $20\log\, [2R_0/(2R_0 - R)]$. Since we no longer have an input and an output pair of terminals this device amplifies equally well in both directions and it has proved so useful that more sophisticated systems using a bridged-T network containing two negative resistances are in service in large numbers.

The negative resistance amplifier can be thought of as a two-terminal device and it can be used to reduce the effects of dissipation in the passive elements of a normal filter. It can also be considered as a four-terminal device in which an impedance Z connected across one pair of terminals produces an apparent impedance $-Z$ at the other pair. This mode of operation leads to the use of the name "negative impedance converter", abbreviated to NIC in the literature. There are three questions to be answered: how do we make an NIC, what are its characteristics, and how do we use it?

When we start to deal with the first question, the question of how to make an NIC, we find that there are two basic kinds of NIC which must be distinguished. Are we to produce a short-circuit stable or an open-circuit stable negative resistance? Some aspects of the way these two types of circuit behave can be studied in my articles on Transistor Inverters (*Wireless World*, Jan., Feb., 1962). If we take Fig. 1 of the first article and redraw it as Fig. 1 here we can copy the result for the resistance R seen at the terminals AB from the January issue of *Wireless World*

$$R = -\frac{1}{k}\left[\frac{Z_1R_2}{Z_2\alpha} + \frac{R_2}{\alpha} + \frac{Z_1(1-\alpha)}{\alpha}\right]$$

If we make $Z_1 = 0$ and $k = 1$ we get

$$R = -R_2/\alpha$$

There are some approximations used in arriving at this result and we can obtain a slightly different form: indeed, we shall. Both are approximately equivalent to $R = -R_2$ as $\alpha \to 1$.

Now in fact we know that $\alpha = \alpha_0/(1 + j\,\omega/\omega_0)$, so that if we take $\alpha_0 \to 1$ we get

$$R = -R_2(1 + j\,\omega/\omega_0)$$

This is an expression which we have already used in determining the trajectory of the working point and it is this form which tells us that the system is short-circuit stable. The trajectories show what happens when a high enough resistance is connected across the terminals AB: the working point flies off looking for a non-linearity to lean against. It finds this, for α must fall as the transistor overloads and as α falls the value of $|R|$ rises until it exceeds the load resistance.

We now modify the circuit of Fig. 1 to eliminate the transformer and to operate as a push-pull system. If we leave out the resistances through which we must feed the bias and standing collector currents we have the circuit shown in Fig. 2. Let us assume that a signal current i_c is flowing in R_2. Most of this, αi_e to be exact, will flow upwards at the collectors of Tr1 and Tr2. A small current $(1 - \alpha)i_e$ will flow in to the base of Tr2 so that the current flowing in at A will be $(1 - \alpha)i_e$ flowing through the capacitance and $-\alpha i_e$ flowing through Tr1. The voltage which produces this current of $(1 - 2\alpha)i_e$ is found by following round the heavy line and if we can ignore the impedance of the capacitances is

$$V_{AB} = i_eR_2 + 2v_{be}$$

We neglect the last term, to get

$$V_{AB} \approx i_eR_2$$

Since this voltage produces a current $(1 - 2\alpha)i_e$ the impedance seen across AB must be

$$R_{AB} = i_eR_2/(1 - 2\alpha)i_e = R_2/(1 - 2\alpha)$$

and as $\alpha \to 1$ this becomes

$$R_{AB} = -R_2$$

The difference between this result and the one previously obtained is the consequence of taking into account the finite base current in the second treatment. The approximation $\alpha \to 1$ makes $(1 - \alpha)i_e \to 0$ and thus produces the same final result.

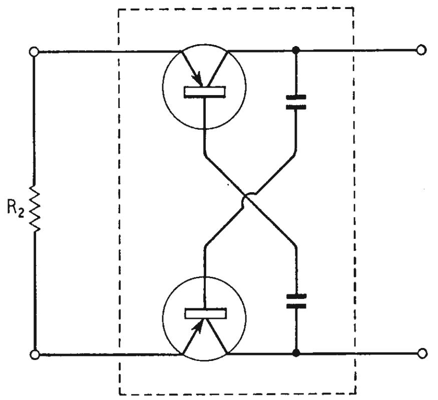


Fig. 3. Rearrangement of Fig. 2, reversed current converter.

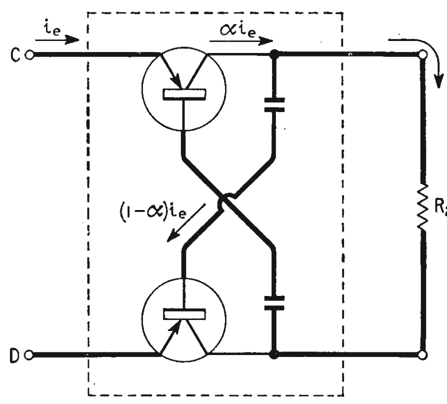


Fig. 4. Reversed voltage converter.

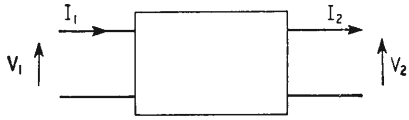


Fig. 5. Block diagram for four terminal network.

When we come to use negative impedance converters for filter design we find that we regard them as networks through which signals are passed. It is therefore convenient to redraw the circuit in the form shown in Fig. 3. The question which immediately arises is what the impedance seen by R_2 will be. If we transfer R_2 to the other end of the network, as in Fig. 4, and consider that a current i_e again flows into the emitter of one transistor we can easily see that the current through R_2 must be αi_e from the collector minus $(1 - \alpha)i_e$ fed to the base of the lower transistor. The net current through R_2 is then $(2\alpha - 1)i_e$ and thus the voltage across R_2 is $(2\alpha - 1)i_eR_2$. The voltage between C and D is obtained by following the heavy line in the diagram and if we now neglect v_{be} we see that the voltage is simply $-(2\alpha - 1)i_eR_2$ for a current i_e and thus the impedance across CD is $R_{CD} = (1 - 2\alpha) \times R_2 \approx -R_2$.

This circuit is open-circuit stable. An alternative derivation of these equations is given with a good deal of circuit detail by Bonner, Garrison amd Kopp in *Bell System Technical Journal* (Nov. 1960, p. 1445).

It will be observed that the circuit of Fig. 4 acts to reverse the voltage applied across CD. Fed for stability from a high impedance source we find that in the block diagram of Fig. 5 we have

$$V_1 = -V_2$$
$$I_1 = I_2$$

The circuit of Fig. 3, twisted round so that the input is on the left, gives us the equation

$$V_1 = V_2$$
$$I_1 = -I_2$$

We see that the two types can be called reversed voltage and reversed current converters.

Practical circuits of this kind commonly use compound pairs of transistors in place of the single units shown so that if, for example, we have $\alpha = 0.98$

(so that $h_{fe} = 50$) we have an effective α for a compound pair of $\alpha_1 + \alpha_2 - \alpha_1\alpha_2 = 0.9996$.

The error in writing $\alpha \approx 1$ is then so small that there is no doubt about the safety of neglecting it: this also carries with it the implication that any changes in the value of α can be ignored.

The two circuits shown in Figs. 3 and 4 are the two basic forms used in negative impedance repeaters where the balanced structure is not a disadvantage. Most frequency selective applications, however, operate in unbalanced, one side earthed, circuits. There would be little merit in eliminating inductances from the filter if we then had to use transformers to get the earth off the centre-point of the active system.

A reversed voltage NIC is shown in Fig. 6 and in rearranged form in Fig. 7. We can see at a glance that the approximate voltage gain from the base of Tr2 to the collector of Tr1 is $(R_2/R_1)(R_S/R_L)$. If this exceeds unity the feedback connection dotted in will produce instability. By taking $R_1 = R_2$ we can say that when $R_L = R_S$ the total resistance must be just zero, so that R_S must see $-R_S = -R_L$. Although we started with the base of Tr2 there is no reason why we should not start with the base of Tr1. Both transistors can therefore profitably be made compound pairs. In this circuit we are using rather a conventional amplifier form and we can regard R_1 as providing local negative feedback. We know that deviations from the ideal negative impedance conversion occur as the transistor α falls with frequency: a conventional video-frequency technique for holding up the gain is to connect a capacitance across R_1 to reduce the negative feedback effect and thus provide an increase in gain to balance the fall due to the frequency dependence of α. This circuit is quoted by Larky (*I.R.E. Trans. Circuit Theory*, Sept. 1957, p. 124) and is due to Linvill (*Proc. I.R.E.*, June 1953, p. 725). A current-inversion circuit due to Larky must now be considered.

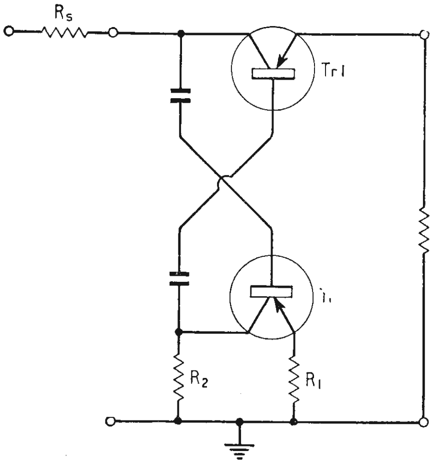


Fig. 6. Reversed voltage unbalanced NIC.

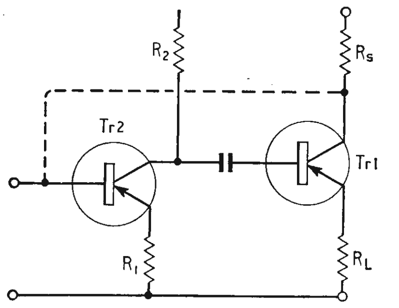


Fig. 7. Rearrangement of Fig. 6.

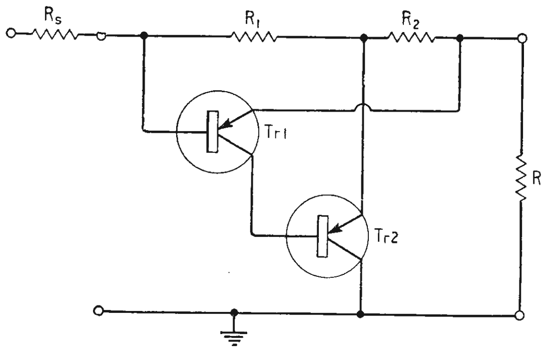


Fig. 8. Reversed current unbalanced NIC.

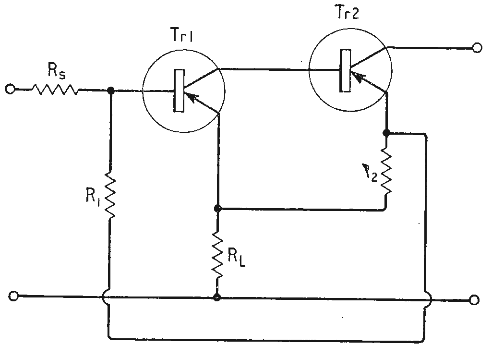


Fig. 9. Rearrangement of Fig. 8.

The circuit is shown as a four-terminal network in Fig. 8 and rearranged to appear as an amplifier in Fig. 9. We see that R_L controls the positive feedback, while R_1 and R_2 control the negative feedback. A detailed analysis of this circuit is given by Larky (*loc cit*). He shows that we need only make Tr2 a compound pair, thus saving one transistor. In addition this circuit has the advantage that the effect of collector capacitance is less than it is in Linvill's circuit and that the harmonic distortion is also less.

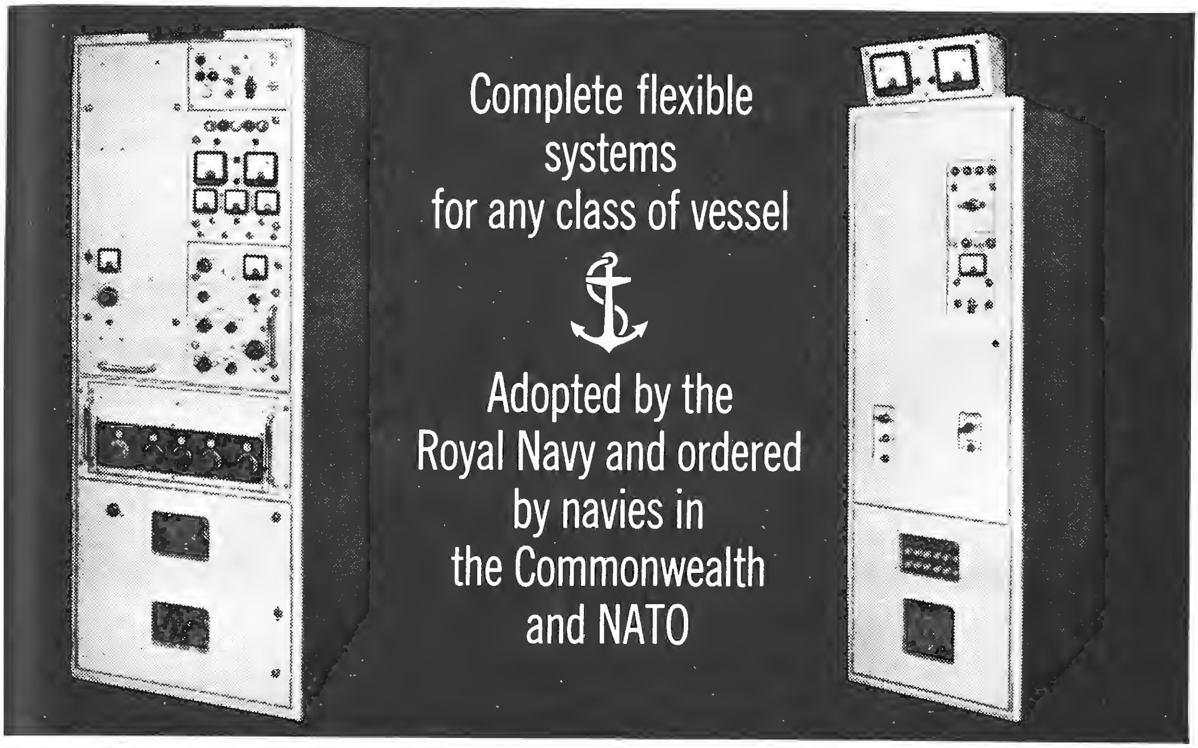
An unbalanced current-inversion type of NIC using complementary transistors has been described by Yanagisawa (*I.R.E. Trans. Circuit Theory*, Sept. 1957, p. 140). The circuit is shown in Fig. 10 with input terminals A B or B′ and output terminals C D or D′. B and D are the common terminals given by Yanagisawa but for a.c. signals it is clear that B′ and D′ can be used and that this gives us a more conventional earthing point. The approximate equations connecting the resistance values for unity conversion ratio are

$R_1 = R_5$, $R_4 = r_e + (1 - \alpha) r_b$, $R_2 = R_3 + r_e$

It is interesting to compare Fig. 10 with Fig. 11,

(*Continued on page 373*)

REVOLUTIONARY NAVAL RADIO TRANSMITTING SYSTEMS



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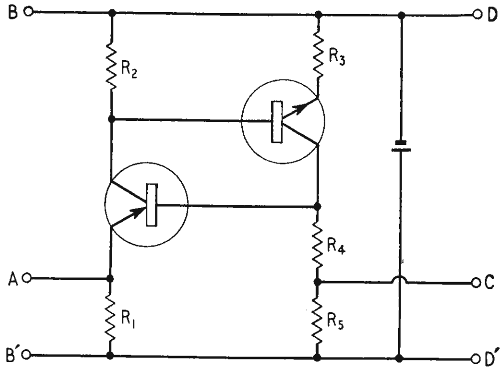


Fig. 10. NIC using p-n-p and n-p-n transistors.

which shows the conventional way of producing a p-n-p-n switch using a pair of transistors. A low impedance across CD′, like a low impedance in the base of a point contact transistor, tends to stabilize the system, and it is seen that this is because it drains away signal which would otherwise reach the base of the p-n-p transistor. The higher the impedance between A and B′ the more negative feedback there will be in the emitter of the p-n-p transistor. Thus the left-hand end is open-circuit stable and the right-hand end is short-circuit stable. It does not seem impossible that this particular circuit should be produced as a single unit.

The next question we must consider is that of the basic characteristics of negative impedance converters. We have already written these down in connection with Fig. 5.

$$\begin{aligned} V_1 &= \mp V_2 \\ I_1 &= \pm I_2 \end{aligned}$$

in which either both upper or both lower signs are to be taken.

We may compare these two equations with the equation for an ideal unity ratio transformer:

$$\begin{aligned} V_1 &= V_2 \\ I_1 &= I_2 \end{aligned}$$

and for an ideal unity ratio gyrator:

$$\begin{aligned} V_1 &= I_2 \\ I_1 &= V_2 \end{aligned}$$

We must add to this collection the ideal negative impedance inverter:

$$\begin{aligned} V_1 &= \pm I_2 \\ I_1 &= \mp V_2 \end{aligned}$$

This has the property that instead of seeing $-R_L$ when we look in at one end with the other terminated in R_L, we see, in general, $-r^2/R_L$, where r is the "ratio" of the device. We can use a negative impedance inverter with a negative impedance

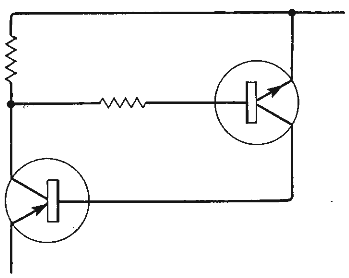


Fig. 11. p-n-p-n switch.

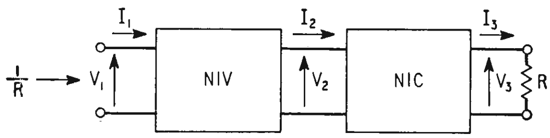


Fig. 12. Gyrator produced by NIV and NIC.

converter to produce a gyrator, for if we have in Fig. 12

$$\begin{aligned} V_1 &= I_2 \\ I_1 &= -V_2 \\ \text{and } V_2 &= -V_3 \\ I_2 &= I_3 \\ \text{we get } V_1 &= +I_3 \\ I_1 &= +V_3 \end{aligned}$$

the form for a gyrator. We can show that a NIV can be produced by using a NIC and three resistances, as in Fig. 13. Two similar NIC's in tandem will give us a unity ratio transformer. The point of this exercise is to show that these new classes of device are different from each other but are all inter-related: the knowledge that there are several closely similar forms serves to remind you to make sure which one you are examining at the moment.

Having established the sort of circuit which can

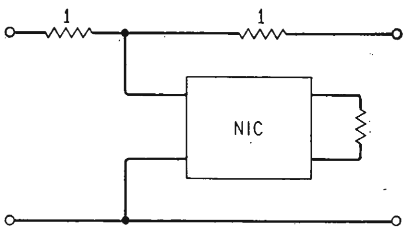


Fig. 13. NIV produced from a NIC.

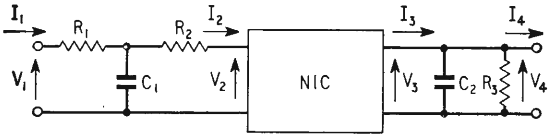


Fig. 14. Network for analysis.

be used and the basic characteristic which will be obtained we can now turn our attention to the use of negative impedance converters in the design of filters. The author knows that whatever he does at this point is wrong. For this purpose there are two readers: one of them has been working full time on this class of problem since 1954 and is surprised that I have neglected the important new work described in his paper, published next week; the other, hot from the calculation of permutations and combinations which leave me dizzy, expresses his hatred of all mathematics. Putting aside the only safe solution, which is to give up altogether and take to science fiction, the answer is to say to the first reader "I know, but they won't read it, they won't even print it". The second reader, if he rejects all mathematics, ought to give up and become a bookmaker's clerk, or go into the Treasury: he is like a one-armed paperhanger, and now one of them will write to the Editor.

There is a fairly easy synthesis given by Shea on

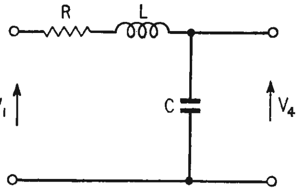


Fig. 15. LC filter equivalent to Fig. 14.

p. 417 of *Transistor Circuit Engineering* (Chapman & Hall). Synthesis is always more difficult than analysis, so what I propose to do is take the structure for granted and analyse its behaviour. This is a how-it-works operation and begs the question of why we chose that network to begin with. The network is shown in Fig. 14. We can write down some equations for this system fairly easily. We have

$$V_1 = (1+j\omega C_1R_1)V_2 + [R_1 + R_2(1+j\omega C_1R_1)]I_2$$
$$I_1 = j\omega C_1V_2 \qquad + (1 + j\omega C_1R_2)I_2$$

for the passive system before the NIC;

$$V_2 = V_3$$
$$I_2 = -I_3$$

for the NIC; and

$$V_3 = V_4$$
$$I_3 = [(1 + j\omega C_2R_3)/R_3]V_4 + I_4$$

for the passive system after the NIC.

Life is made a little easier by the fact that $I_4 = 0$, because there is nowhere for the current to go. A certain amount of elimination and we arrive at the result

$$V_1/V_4 = (1 + j\omega C_1R_1) - [R_1 + R_2(1 + j\omega C_1R_1)]\,[1 + j\omega C_2R_3]/R_3$$

which when $\omega = 0$ becomes

$$\frac{V_1}{V_4} = 1 - \frac{R_1 + R_2}{R_3}$$

If we want $|V_1/V_4| = 1$ at $\omega = 0$ we can either take $(R_1 + R_2) = 0$, $R_3 = \infty$ or take $(R_1 + R_2)/R_3 = 2$. This last form is the only one which leaves us with the whole structure and it makes $V_1/V_4 = -1$, showing a phase reversal.

Then we have

$$\frac{V_1}{V_4} = (1+j\omega C_1R_1) - \frac{R_2}{R_3}\left(2\frac{R_3}{R_2} + j\omega C_1R_1\right)\left(1 + j\omega C_2R_3\right)$$
$$= -1 + j\omega\left(C_1R_1 - \frac{C_1R_1R_2}{R_3} - 2C_2R_3\right) + \omega^2C_1C_2R_1R_2$$
$$= -\left\{1 + j\omega\left[2C_2R_3 + C_1R_1\left(\frac{R_2}{R_3} - 1\right)\right] - \omega^2C_1C_2R_1R_2\right\}$$

Let us now take a look at the ordinary second-order filter shown in Fig. 15. For this we have

$$\frac{V_1}{V_4} = 1 + j\omega CR - \omega^2LC$$

and $$\left|\frac{V_1}{V_4}\right|^2 = 1 + \omega^2(C^2R^2 - 2LC) + \omega^4L^2C^2$$

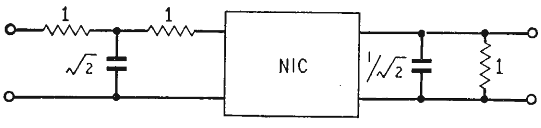


Fig. 16. Network with Butterworth response.

The 3dB point with a Butterworth response is at

$$\omega_0^2\, LC = 1$$

and the shape requires that

$$R^2 = 2L/C$$

We can get exactly the same response shape from our network of Fig. 14 by arranging that

$$C_1\, C_2\, R_1\, R_2 = LC$$

and $$2C_2\, R_3 + C_1\, R_1\left(\frac{R_2}{R_3} - 1\right) = CR$$

We now have five elements in place of three, so that we cannot just solve the equations. This shows up in the synthesis procedure given by Shea, in which two positive numbers must be chosen arbitrarily. Let us take as the design cut-off a frequency $\omega_0 = 1$ and for the circuit of Fig. 15, $R = \sqrt{2}$. Then $L = 1$, $C = 1$, $R = \sqrt{2}$. This gives us

$$C_1C_2R_1R_2 = 1$$

and $$2C_2R_3 + C_1R_1\left(\frac{R_2}{R_3} - 1\right) = \sqrt{2}$$

Now let us choose $R_2 = R_3$, so that since $(R_1 + R_2)/R_3 = 2$, we also have $R_1 = R_3$. Let $R_1 = R_3 = 1$; then we find that $C_1\, C_2 = 1$. Since $R_2/R_3 - 1 = 0$, we have $2\, C_2R_3 = \sqrt{2}$. Thus $C_2 = 1/\sqrt{2}$, and $C = \sqrt{2}$.

The final result is shown in Fig. 16. Shea's synthesis is based on a somewhat different response shape, $(1 + j\omega - \omega^2)$, so that he requires $|V_1/V_4|^2 = (1 - \omega^2 + \omega^4)$, a response with a hump, and obtains rather different values.

We could have chosen $C_1 = C_2$, though the calculation is slightly less convenient. This, however, offers us a system in which by switching pairs of identical capacitors we can move the cut-off frequency without altering the shape of the response. We can provide as many steps as we wish without any of the problems which arise when tapped inductors are used.

Physically we may regard the system we have thus obtained as two first order RC filters in tandem, giving us the wanted 12dB/octave cut-off slope, with feedback through the amplifier to sharpen up the characteristic round the knee. By using more elaborate networks, but still only RC networks, on either side of the NIC we can produce higher order filters although the characteristics of the NIC will become more and more critical. Since we have a phase reversal we can also provide a path round the first network and the NIC which will produce at some frequency a current I_3' exactly equal to I_3, so that at this particular frequency we must have $V_4 = 0$. In this way we can get the attenuation peak we associate with a conventional m-derived filter.

At the present time any filter of this sort must be designed by direct synthesis or by a hodge-podge method of the sort used in this article. It cannot be very long, however, before somebody settles down to the collation of all the existing designs and gives us tabulated structures and values so that we can build this sort of structure in the same way as we build the simple LC filter.

Perhaps it has already been done; perhaps I shall do it myself. In any event, to use the tables you will still need to know a little about how the filter works.

FUNDAMENTALS OF FEEDBACK DESIGN

8.—THE STABILITY MARGIN

By G. EDWIN

THE measurement and interpretation of the loop gain and loop phase characteristics over the necessarily very wide range of frequencies demands time and skill. In production testing the advantage of a simplified method will be obvious and even in the construction of a single amplifier, whether it be an amplifier alone, as it were, or the prototype for future production, a single easily measured criterion is of great value. The stability margin, described by Duerdoth* gives us such a criterion.

To understand the meaning of the stability margin we must turn to the Nyquist diagram. As before we shall concern ourselves with only one end of the frequency response although we shall remember that whatever we say about one end applies equally to the other. The Nyquist diagram in Fig. 40 has been drawn from the responses already used in our June issue (Figs. 29 and 30) and may be regarded as fairly typical except that it shows a sharp corner at $\omega=1$ due to the fact that it is based on a straight line approximation. The vector OP represents the quality $\mu\beta$ and it will be seen that the trace does not encircle the critical point (1, 0) at R. Indeed, at A, where $|\mu\beta|=1$, there is a margin of 30° and at B, where the phase angle has come round through 180°, there is a gain margin of 6dB.

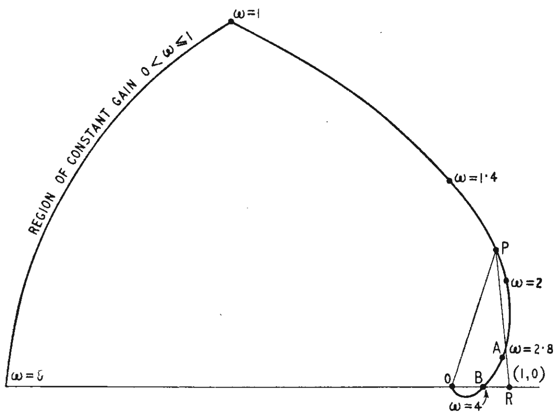


Fig. 40. $\mu\beta$ diagram of a system having two $[1 + j\Omega]$ terms with characteristic frequencies at $\omega = 1$ and a third at $\omega = 4$ (see amplitude and phase responses of Figs. 29 and 30 of part 6 of this series on p. 282 of our June issue).

We know, however, that although we draw a neat line on this diagram this is really rather meaningless. Component tolerances and valve gain variations can alter both the scale of the diagram and its shape: we should do better to draw our curve with a poster pen. One way in which we can take some account of these variations is by determining how near to R the curve passes. In a way this is what we do when we examine the two points A and B, the points defining phase and gain margins. Somewhere between these two, in Fig. 40, is the point of nearest approach to the point R. This diagram is plotted in the $\mu\beta$ plane, of course, which explains why the interesting region is to the right of the origin.

Vectorially, OP=OR+RP, so that the vector RP is equal to $\mu\beta-1$. We want to know when this is the shortest distance, so we want to know the minimum of $|\mu\beta-1|$. Let us write the gain with feedback as μ_f, and we have

$\mu_f=\mu/(1-\mu\beta)$
so that $\mu/\mu_f=1-\mu\beta$
and thus $|\mu/\mu_f|=|1-\mu\beta|=-|\mu\beta-1|$

The length RP is thus equal to the ratio of the gain without feedback to the gain with feedback, apart from a minus sign. As we are working in decibels we use $20\log|\mu_f/\mu|$ and it is this which Duerdoth has called the stability margin. Away from the critical region this is a negative quality and for this particular system is, in fact, -18dB. Above $\omega=1$ the stability margin changes rapidly and its behaviour is shown in Fig. 41 for the 18dB feedback case corresponding to curve I of Fig. 29 (June issue). The important point is the peak of $+7.5$dB which corresponds fairly closely to the response peak. This is the nearest approach to instability and represents, in fact, a condition which is not quite acceptable. The safe limit which was recommended by Duerdoth was $+6$dB.

The great advantage of the stability margin in practice is that it is extremely easy to measure. A signal is applied to one input of the system and the feedback disconnected: normally the output will rise by the feedback factor. As the frequency is moved towards the edge of the band, and we are talking in this example of the upper edge, though the same effect is seen at the lower edge, the rise in output when the feedback path is broken falls rapidly. We are interested in the region where disconnecting the feedback produces a small increase in gain and it is very quickly identified. Since the gain change

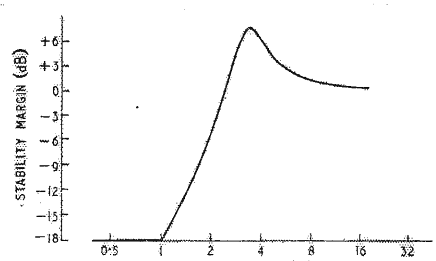


Fig 41. Plot of Stability margin corresponding to Fig. 40 and 18dB feedback in Fig. 29 (curve I).

* *Proc. I.E.E., Part III*, Vol 97, p. 138 (May 1950).

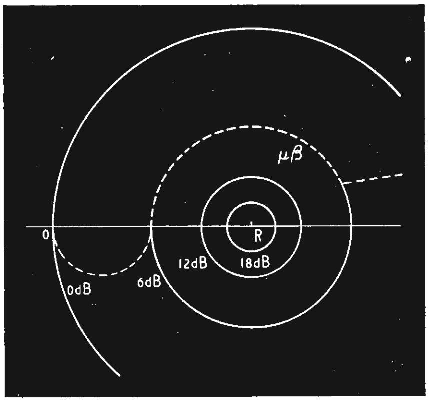


Fig. 42. Circles of constant stability margin in μβ diagram and ideal μβ characteristic in neighbourhood of critical point, showing 6dB stability margin, 6dB gain margin and 30° phase margin.

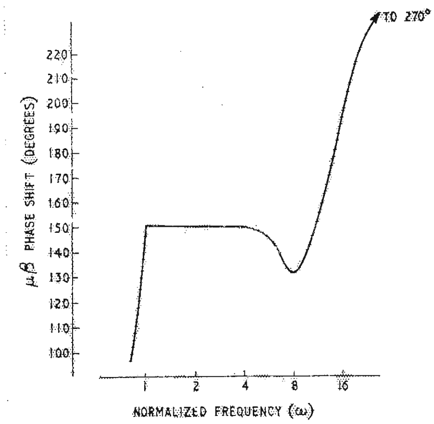


Fig. 44. Phase response showing dip in angle associated with step in amplitude response.

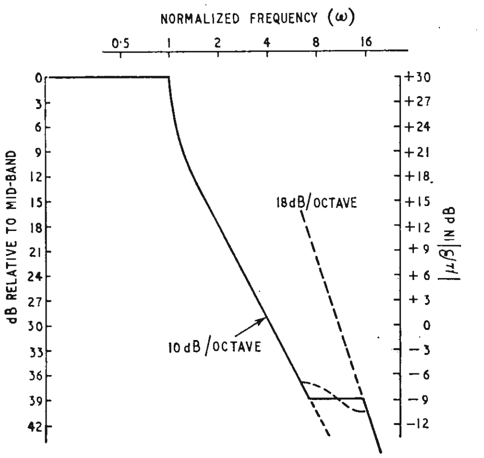


Fig. 43. Amplitude response showing stabilizing step.

should be only a few decibels we can set the input level to give us a convenient reading on the output meter and read off the stability margin directly as the change of reading. In this way it is a very quick operation to locate the peak. We now know if the amplifier is safe and we also know, when it is unsafe, the exact location of the danger point in order that we may calculate a step circuit to increase our margin.

The only defect which this system has is a terminological one. The normal designs have stability margins of 0 to +6dB but the higher the stability margin the less stable the amplifier will be. It would be preferable if another name could be used for this quantity, a name such as "ringing coefficient", which would remind the user that a high stability margin is not desirable.

This result is very much the same as we should obtain by square-wave excitation and study of the ringing at high frequencies. It is, however, equally applicable to the low frequency response. Furthermore, in routine testing it does not require any judgement. The answer is a number which can be specified to be less than whatever the designer thinks fit. Either it is, or it is not: there is no question of deciding that it looks just about all right.

Lines of constant stability margin are obviously circles centred on R in the $\mu\beta$ plane: the 0dB stability margin circle, with unit radius, passes through the origin and the radius of the 6dB circle is $\frac{1}{2}$, of the 12dB circle $\frac{1}{4}$ and so on. Remember that $\mu\beta$ is not plotted on a logarithmic scale. These circles are shown in Fig. 42. It would be very desirable to provide a $\mu\beta$ characteristic, of the form shown in Fig. 42, which stays high and outside this diagram until the angle is perhaps 150° or more and then runs in towards the origin, dodging round the point R along the 6dB stability margin circle until it crosses the axis and then making its own way back to the origin.

The separate amplitude and phase responses which correspond to this sort of behaviour are generally similar in form to those given by Bode, which are of the type indicated in Figs. 43 and 44. In Bode's treatment a flat step is used in the amplitude response to produce the dip in the phase characteristic: in the response derived from the stability margin graph the low slope region will have something of the form indicated by the chain line. Since the phase angle depends on the weighted average slope, the local detail is not very material.

It has been pointed out, however, that gain is more variable than phase. At the characteristic frequency of a $[1+j\Omega]$ circuit, an error of 10% in either the reactance or the resistance will only produce 3° error in the phase, so that the six basic elements of a three-stage amplifier cannot swing the phase more than 20°. Gain variations, however, can easily be more than 6dB for a three-stage amplifier. To allow for this an ideal $\mu\beta$ characteristic might have the form shown in Fig. 45, which will allow a gain increase of about 10dB before instability occurs, has a phase margin of

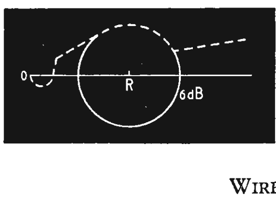


Fig. 45 Form of μβ characteristic which allows for more gain tolerance than Fig. 42.

30° and hugs the 6dB stability margin in a way which leads to the most efficient use of the available gain-bandwidth product. This sort of design is rather sophisticated, of course, and the reader who is interested is recommended to refer to Duerdoth's paper (*loc. cit.*).

It is rarely convenient to use the Nyquist diagram because it can only be constructed by first constructing the amplitude and phase diagram and then turning to the anti-log tables. We need some way of determining the stability margin without passing through the Nyquist diagram.

Let us write for the stability margin

$$\sigma = 20\log(\mu_t/\mu)$$

and thus

$$\sigma = 20\log[1/(1-\mu\beta)]$$

$$= 20\log\left[\frac{1}{\mu\beta}\Big/\left(\frac{1}{\mu\beta}-1\right)\right]$$

$$= 20\log\left[\frac{-1}{\mu\beta}\Big/\left(1-\frac{1}{\mu\beta}\right)\right]$$

We have seen previously that the mu-beta effect calculator enables us to determine the quantity $-\mu\beta/(1-\mu\beta)$. If we make use of the calculator in exactly the same way but find the $\mu\beta$-effect corresponding to $-(\mu\beta$ in dB) instead of to ($\mu\beta$ in dB) we can get the values of the stability margin directly from the amplitude and phase plots. This result was given by Jefferson in *Proc. I.R.E.*, Vol. 39, p. 1571 (Dec. 1951), from which the table below is taken.

Normal Gain

ω	$-\lvert\mu\beta\rvert$dB	$\theta°$	σdB
1.5	−17	37	−16
2	−13	44	−11.6
4	−5	56	−3.4
10	4	57	+1.5
20	15	23	+1.6

Gain 12dB up

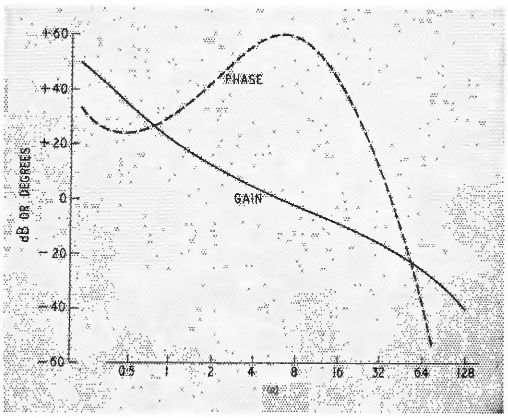
ω	$-\lvert\mu\beta\rvert$dB	$\theta°$	σdB
1.5	−29	37	−29
2	−25	44	−25
4	−17	56	−16.5
10	−8	57	−6.6
20	+3	23	+7.1

This table relates to an amplifier discussed in the paper by Felker (*Proc. I.R.E.*, Vol. 37, p. 1204, Oct. 1949) which describes the use of the calculator. The two sets of figures correspond to an amplifier of normal gain and one having 12dB more gain than standard. With this rather high gain the stability margin is +7dB, a figure which might be regarded as tolerable for this top-limit case. The characteristics of the amplifier are plotted out in Figs. 46 and 47.

In practice the stability margin can be worked out at the same time as the frequency response or, more commonly, in the preliminary stages of the design only the stability margin will be found by determining the few points necessary in the critical region. The design treatment then models exactly the method which will later be used for testing the system.

An interesting special case is the rather synthetic example shown in Figs. 48, 49 and 50. In this the stability margin has been made 0dB and the sort of response shape involved is shown in Figs. 49 and 50. The gain with feedback applied is asymptotic to the gain without feedback curve and the difference between them is in fact a combination of a small centring error in the calculator and drawing errors. The critical student will note that this must be a single stage amplifier since the phase is asymptotic to 90°. The low frequency end of an amplifier with only one a.c. coupling could have this form. There are

Fig. 46. Amplitude and phase response without feedback of amplifier discussed by Felker in Proc. I.R.E., Vol. 37, p. 1204, Oct. 1949 (based on Fig. 2 of this article).



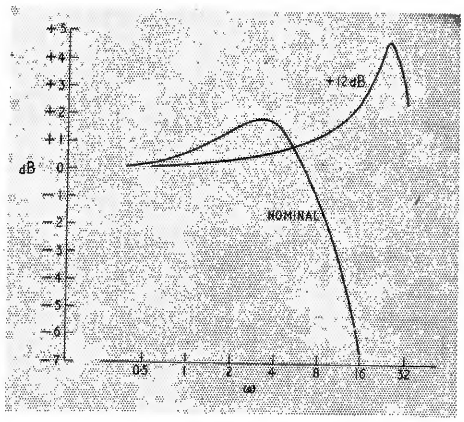


Fig. 47. Amplitude response with feedback for the amplifier of Fig. 46 for the two cases of normal gain without feedback and gain increased by 12dB (based on Fig. 3 op cit).

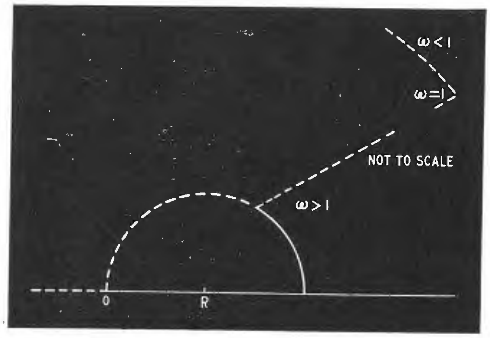
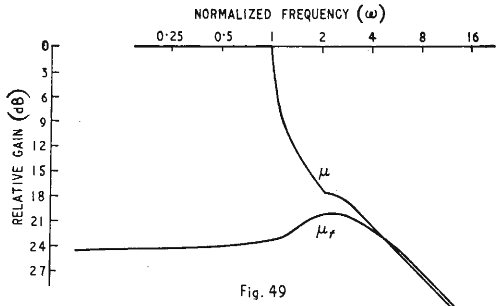
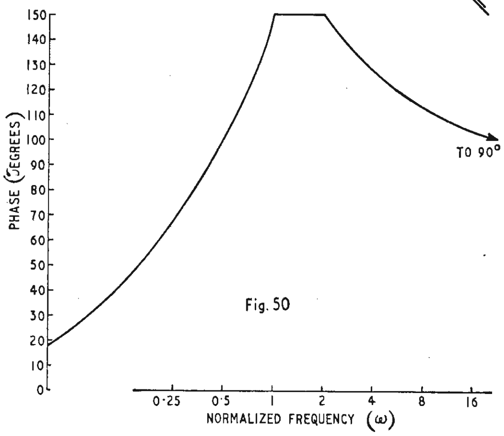


Fig. 48. $\mu\beta$ characteristic for zero stability margin, a very conservative design.





Figs. 49, 50. Amplitude and phase response of a rather artificial system which will give the zero stability margin characteristic shown in Fig. 48, and which shows the monotonic approach of the gain with feedback curve to the gain without feedback.

interesting possibilities in the use of this as a design method when very carefully controlled roll-off shapes are needed.

Although it has been assumed that the reader is using a mu-beta effect calculator, the form of the stability margin makes it possible to get results without too much trouble on a slide rule. We can write again.

$\sigma = 20\log 1/(1-\mu\beta)$
$= -20\log(1-\mu\beta)$

We know $20\log|\mu\beta|$ and the phase angle θ, so that we can easily get $|\mu\beta|$, and thus find $|\mu\beta|\cos\theta$ and $|\mu\beta|\sin\theta$. Then we need to tabulate $1-|\mu\beta|\cos\theta$ and $|\mu\beta|\sin\theta$. These we square, add, and without taking the square root, find simply $-10\log[(1-|\mu\beta|\cos\theta)^2 + (|\mu\beta|\sin\theta)^2]$. This is the stability margin. We can plot this as a distance above the μ characteristic and we have our response. Naturally this is more tedious than using the calculator: if not, why should we bother to get a calculator.

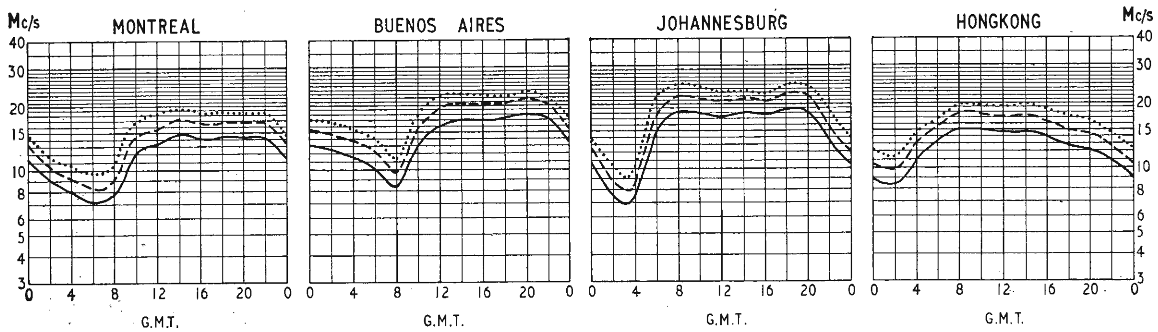
We have now covered the basic ground connected with the design of single loop negative feedback systems so that it is a convenient point to look back and see how essentially simple the procedure can be. The amplifier circuit is sketched out and we make sure that it will give the output power we need and that there is the gain required, the final overall gain plus the amount of feedback we have chosen. Piece by piece we build up the amplitude and phase characteristic, modifying circuit elements as we go to preserve reasonable phase and amplitude margins. If necessary we include simple step circuits to bring down the amplitude by increasing the phase shift in a safe region while allowing it to drop back again as the danger region is approached.

Once a probable solution is reached we can calculate the amplitude and phase characteristic more exactly than with the straight line approximation. Now we use the mu-beta calculator, to find the response with feedback connected and the stability margin. Final corrections can be made and we are ready to begin construction.

There remains, however, what may be called fringe topics although some of them are vital to the subject. The most directly concerned are the question of distortion in systems with negative feedback and the effect of feedback on the terminal impedances of amplifiers. This leads us to consider the use of positive feedback, from which we must go on to the special application of positive feedback in the production of well-defined negative impedances.

SHORT-WAVE CONDITIONS

Prediction for August



THE full-line curves indicate the highest frequencies likely to be usable at any time of the day or night for reliable communications over four long-distance paths from this country during August.

Broken-line curves give the highest frequencies that will sustain a partial service throughout the same period.

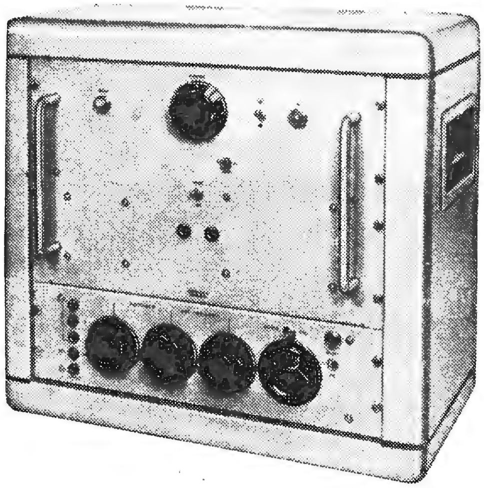
•••••••••• FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE FOR 25% OF THE TOTAL TIME

— — — PREDICTED MEDIAN STANDARD MAXIMUM USABLE FREQUENCY

——— FREQUENCY BELOW WHICH COMMUNICATION SHOULD BE POSSIBLE ON ALL UNDISTURBED DAYS

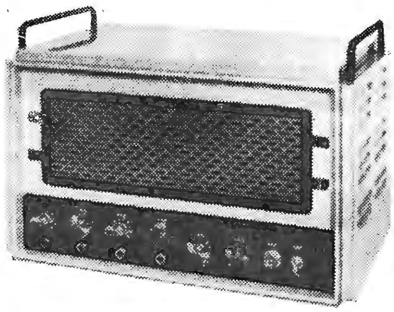
Vortexion quality equipment

120/200 WATT AMPLIFIER



Will deliver 120 watts continuous signal and over 200 watts peak Audio. It is completely stable with any type of load and may be used to drive motors or other devices to over 120 watts at frequencies from 20,000 down to 30 cps in standard form or other frequencies to order. The distortion is less than 0.2% and the noise level —95 dB. A floating series parallel output is provided for 100-120 V. or 200-250 V. and this cool running amplifier occupies 12¼ inches of standard rack space by 11 inches deep. Weight 60lb.

30/50 WATT AMPLIFIER



Gives 30 watts continuous signal and 50 watts peak Audio. With voice coil feedback distortion is under 0.1% and when arranged for tertiary feedback and 100 volt line it is under 0.15%. The hum and noise is better than —85 dB referred to 30 watt.

It is available in our standard steel case with Baxandall tone controls and up to 4 mixed inputs, which may be balanced line 30 ohm microphones or equalised P.U.s to choice.

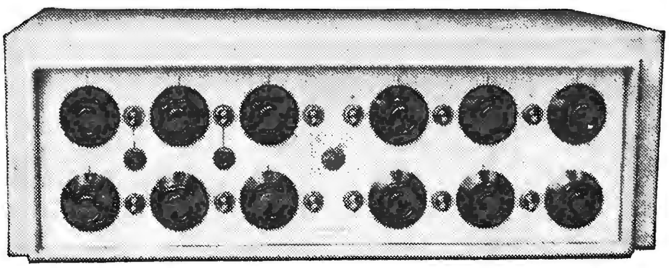
ELECTRONIC MIXER/AMPLIFIER

This high fidelity 10/15 watt Ultra Linear Amplifier has a built-in mixer and Baxandall tone controls. The standard model has 4 inputs, two for balanced 30 ohm microphones, one for pick-up C.C.I.R. compensated and one for tape or radio input. Alternative or additional inputs are available to special order. A feed direct out from the mixer is standard and output impedance of 4-8-16 ohms or 100 volt line are to choice. All inputs and outputs are at the rear and it has been designed for cool continuous operation either on 19 x 7in. rack panel form or in standard ventilated steel case.

Size 18 x 7½ x 9½in. deep.

Price of standard model **£49.**

12-WAY ELECTRONIC MIXER



The 12-way electronic mixer has facilities for mixing 12 balanced line microphones. Each of the 12 lines has its own potted mumetal shielded microphone transformer and input valve, each control is hermetically sealed. Muting switches are normally fitted on each channel and the unit is fed from its own mumetal shielded mains transformer and metal rectifier.

Also 3-way mixers and Peak Programme Meters. 4-way mixers and 2 x 5-way stereo mixers with outputs for echo chambers, etc. Details on request.

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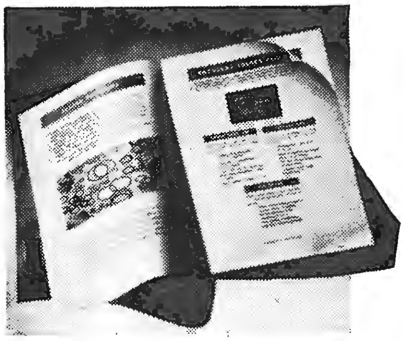

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Excitations and Responses

By "CATHODE RAY"

IN the last four issues we have been looking at the basic theory of circuits from various viewpoints, and I think it is about time to put these views together and try to construct a whole picture. Let's see what we have.

Real circuits, of course, are composed of "devices," which may be quite complicated electrically. So for purposes of study and calculation they are usually replaced by (as nearly as possible) equivalent theoretical circuits, made up of lumps of resistance, inductance and capacitance (R, L and C) arranged in series or parallel or both. To make a circuit work there must at least be one source of voltage or current. A voltage source will drive current through the circuit and a current source will give rise to voltages between points in it. In more general terms, excitations cause responses. Circuit theory is concerned mainly with finding the ratio of response to excitation.

If the response to be considered is between the same pair of terminals as the excitation, this ratio is comprehensively called the immitance—a composite word that can be used for either an impedance (ratio of voltage to current) or an admittance (ratio of current to voltage). If the response is elsewhere than at the excitation terminals, the ratio is called a transfer function. In this case the ratio can be output current to input current (current gain) or output voltage to input voltage (voltage gain) or current to voltage or voltage to current. If input and output are both current or both voltage the magnitude of the ratio is just a number; otherwise it has the dimensions of an immitance. I say *magnitude* because both immitances and transfer functions are in general "complex"—they have phase as well as magnitude.

We have been confining our scope—and will keep on doing so—to linear circuits, so please don't bring in any rectifiers or saturable reactors. And we began by confining our scope to sinusoidal (sine-wave) excitation—and therefore, in our linear circuits, sinusoidal responses—but latterly began to escape a little from this restriction. Before we're finished I hope to widen the bounds still more.

If you have been thinking that the elaborateness of the mathematical apparatus has been absurdly out of proportion to the extreme simplicity of the problems on which I have brought it to bear, you were right. I purposely chose the simplest possible examples to show the methods in principle, unobscured by detail. To review these methods we are going to continue with the example in use towards the end of last month's instalment (Fig. 1 here), except that to make it a little more general I will not fix any particular ratio of R to ωL.

I assumed familiarity with reactance and its calculation, whereby the method of finding i, the response to a *sinusoidal* excitation v, would be something like this: the magnitude of the reactance of L is found by multiplying it by ω $(=2\pi f)$; but the result can't just be added to R to make up the impedance, because it is 90° different in phase. This is indicated by j, signifying 90° anticlockwise rotation on a vector diagram. The total impedance $\boldsymbol{Z}$ is therefore $R + j\omega L$, and from the geometry of the diagram, and Pythagoras, the magnitude Z is $\sqrt{(R^2 + \omega^2 L^2)}$. This is not the ratio v/i, because those are instantaneous values; it is (more conveniently) V/I, the capital letters denoting peak or r.m.s. values.

My guess last month was that a lot of people who are quite capable of applying these rules to any reasonable circuit, and who may even know the elements of the differential calculus, are unaware that the procedure they follow is derived from the solution of a differential equation. Every a.c. circuit involves a differential equation, but by making certain assumptions (e.g., sinusoidal excitation) one can use a stock solution without necessarily knowing where it came from.

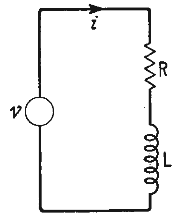


Fig. 1. This simple circuit continues to provide food for a lot of thought.

The equation is formed by applying Kirchhoff; equating all the voltages around a circuit—or all the currents entering a junction—to zero. In this case we have three voltages: the e.m.f. v, and, opposing it, a voltage proportional to R and the current through it (Ri) and one proportional to L and the *rate of change* of current through it (L di/dt):

$$v - \left(Ri + L\frac{di}{dt}\right) = 0 \qquad .. \qquad (1)$$

This goes for *any* form of v or i; not just sinusoidal. Solution is facilitated, however, if we assume it is some kind of exponential function, because they have the unique feature that their rates of change are proportional to themselves; e.g.

$$\frac{d}{dt}e^{at} = ae^{at} \qquad .. \qquad .. \qquad (2)$$

Note that the constant multiplier (a) of the variable (in our examples, t) comes out as a multiplier of the whole exponential function. In the opposite process of integration it comes out as a divisor.

One assumes therefore that i can be expressed as Ae^{at}, where A and a are constants to be found. Substituting in (1) and using (2) results in

$$i = Ae^{at} = \frac{v}{R + aL} \qquad .. \qquad .. \qquad (3)$$

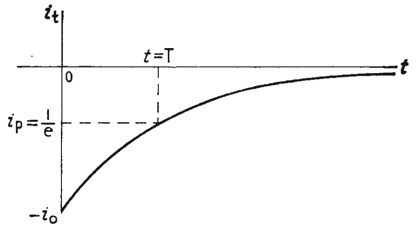


Fig. 2. Exponential die-away curve representing the form of transient response in a circuit of the Fig. 1 type.

As explained last month, a sinusoidal v can be written not only in the familiar form V $\cos\omega t$' where V is the peak value, but in the less obvious one $Ve^{j\omega t}$. The latter, being an exponential function, entitles it to a place in (3). Making the substitution shows that $A = \dfrac{V}{R + aL}$ and $a = j\omega$, so

$$i = \frac{V}{R + j\omega L} e^{j\omega t} = Ie^{j\omega t} \quad .. \ (4)$$

where I is the peak value of what turns out to be a sinusoidal current. We can now divide (4) by $e^{j\omega t}$, eliminating all trace of this occult means of arriving at the familiar relationship

$$I = \frac{V}{R + j\omega L}$$

where the form of simple Ohm's law is preserved, with peak values of current and voltage (or r.m.s., since both sides can be multiplied by $1/\sqrt{2}$ or any other factor) and the idea of resistance is extended into impedance. Or $1/(R + j\omega L)$—which is the admittance—can be regarded as the circuit function by which the excitation V has to be multiplied to find the response, I.

We saw that to be strictly correct we can apply this solution only to a.c. that has been going on with perfect regularity for an infinitely long time. In reality it has to begin, so we considered what happens when it is switched on. As our circuit was assumed linear, we can find the transient switching-on current separately and then add it to the regular one to find the total. The equation for this transient is the same except that $v = 0$, and we assume the transient current (call it i_t) is $A_t e^{a_t t}$:

$$A_t e^{a_t t} (R + a_t L) = 0 \quad .. \quad (5)$$

This would be true (obviously!) if the current were zero, representing the remote chance of the circuit being switched on at a precise moment when i would have been zero even if it had been switched on all the time. In general, however, the "regular" current at the switching-on time ($t = 0$) would be something we will call i_o, and, since current cannot be made to flow in an inductive circuit instantaneously, i_t at this moment must be equal and opposite to i_o, to make the total zero. Because any number to the power 0 is 1, when $t=0$ it makes $i_t = A_t$, so $A_t = -i_o$. And to make (5) true when current is flowing, $a_t L$ must equal $-R$, or $a_t = -R/L = -1/T$, where T is the time constant of the circuit. Thus

$$i_t = -i_o e^{-t/T} \quad .. \quad .. \quad (6)$$

This is plotted as Fig. 2, and again here is something that must be familiar to many who are blissfully ignorant of the differential equation from which it sprang. The point that is always brought out even in the rule-of-thumb course is that when the time after the start is equal to T the amplitude has died away to 37% of its original value (or has built up to within 37% of its final value, if the initial value is taken as the zero level). This follows from (6) because $e^{-1} = 1/e = 0.37$.

Note that this transient current depends only on the circuit's time constant and on the initial current *however caused.* As we saw last month it could just as well have been caused by switching off a battery as by switching on a sinusoidal e.m.f. So the transient is often called the natural or free response of the circuit. By contrast, the form of the "regular" current is the same as that of the excitation, only its magnitude and phase being affected by the circuit; so it is known as the forced response.

Given data on the free response—say Fig. 2 with a numerical time scale—one could find the time constant and hence work back to the circuit that must have caused it, or at least an equivalent circuit. Similarly a graph or other information on the forced response would reveal the form of the excitation. But what if we were to be given the total response?

Since the transient response in our example is negligible after a few cycles, the forced response is the current that persists thereafter, and the transient can easily be found by subtracting that from the first few cycles. So the answer seems obvious, that we could deduce the type of circuit and the excitation. In principle, at least, however, the obvious is wrong. Are we justified in assuming that the short-lived response is the free one and the continuous one the forced? The forced response to a voltage of the die-away form would be as in Fig. 2. And the free response, to *any* excitation, of a circuit containing only L and C, is a continuous sine wave!

Any readers hardy enough to have been coming up for my punishment month after month may be reckoning that I have forgotten all about the poles and zeros which were featured in parts 2 and 3 (we are now in 5), or else that I have abandoned them to their fate. Now, however, with the audience reeling

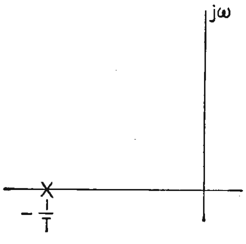


Fig. 3. Pole-zero diagram (actually it contains no zeros and only one pole) for Fig. 1.

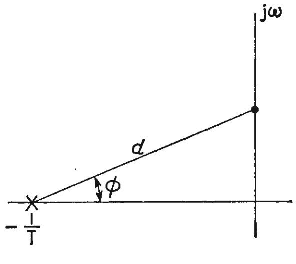


Fig. 4. How to use Fig. 3 to find the magnitude and phase of the response of Fig. 1 to sinusoidal excitation of any frequency.

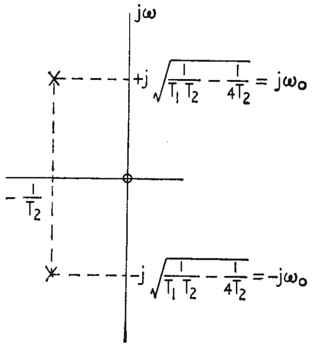


Fig. 5. Pole - zero diagram for a RLC circuit in which R is relatively fairly small.

from the shock of seeing that, so far as can be judged from results, circuits and sources are indistinguishable, is the precise moment poles and zeros are scheduled to make what I hope will be a dramatic reappearance.

Regretfully I must allow the impact of this reappearance to be somewhat blunted by my fear that even readers who have been with us for the last three months will not have memories so retentive that they will instantly grasp the significance of the very simple Fig. 3 and confidently declare it to be the pole-zero diagram corresponding to Fig. 1. On the other hand, complete newcomers will hardly find themselves without any clue, for $j\omega$ and $-1/T$ have both figured in the present discussion. So I hope the action will be held up only so long as is needed for a hurried explanation that the × marks a " pole ", where the frequency is such as to make the circuit function (admittance in this case) infinitely large. " Real " frequency is measured along the " imaginary " axis, as $j\omega$. At the first exposition I had to ask you to accept this apparently illogical arrangement as just one of those things. But since we became familiar with $e^{j\omega t}$ as a practical representation of the ordinary sine wave it begins to look more reasonable. Imaginary frequency (yes; I know!) would by the same logic have to be $j\,(j\omega)$, $= -\omega$. The current-response/voltage-excitation ratio, or circuit function for Fig. 1, as we have already established, is the admittance $1/(R+j\omega L)$, $= 1/R(1+j\omega T)$. It becomes infinitely large at the frequency which makes the denominator zero; i.e., when $\omega = j/T$, making $j\omega = -1/T$ and $j\omega T = -1$. Hence the location of the pole.

The magnitude of the circuit function at any real frequency is obtained, according to the rules we established in the May issue, by locating that frequency on the $j\omega$ scale and drawing lines from that point to all poles and zeros. In this case we have only one pole, so Fig. 4 shows the construction. The magnitude is given by a scale factor (1/R in this case) multiplied by the lengths of the lines to zeros, divided by the lengths of the lines to poles, each length having been divided by the appropriate 1/T. In our example this calculation is $1/R \div (d \div 1/T) = 1/RdT$. This is obviously 1/R at zero frequency, as one can also see from Fig. 1; and it falls as the frequency rises. The phase angle is equal to the sum of the angles of the lines with reference to the horizontal, angles at zeros being reckoned positive and those at poles negative. So here it is $-\phi$. The impedance, being the reciprocal of the admittance, is $RdT\underline{/+\phi}$ in this example.

The one new point that emerges this month is that the pole location, $-1/T$, is the same as the coefficient of t in the exponential form of the transient current, $e^{-t/T}$. This is so important that I suggest (with all respect) that you read it again, slowly. So far we have used pole-zero diagrams for finding only the forced response, not the free response. But now we see that the form of the free response depends solely on the circuit in which it occurs, and in particular the time constants of the circuit. (In more complicated ones there are several time constants, each locating a pole or a zero). So here is a new significance to attach to poles and zeros—they define the free components of the response.

In our example the free response (Fig. 2) has no discernible frequency; it is a simple die-away exponential curve. So the fact that its pole is located on the axis dedicated to imaginary frequency begins to make some kind of sense.

Our previous applications of pole-zero diagrams have been confined to circuits with sinusoidal sources, having real frequency—paradoxically measured along the imaginary ($j\omega$) axis. But the $j\omega$ now rings a bell, for this month we have found it to be the coefficient of t in the exponential form of a sinusoidal excitation, $e^{j\omega t}$. So it forms the perfect counterpart of the real coefficient $-1/T$ denoting an imaginary frequency.

We have noted, too, that the forced response takes the *form* of the excitation, but its *magnitude* and *phase* are determined by the circuit. So here is the logic of the rules used for determining the forced response by measuring lines drawn from the poles and zeros (representing the circuit) to a point on the $j\omega$ or real-frequency axis (representing a sinusoidal source). Both are involved.

Two months ago, when we examined a second-order circuit, comprising R, L and C, we found that it had one zero (because there was a $j\omega$ factor in the numerator) and two poles (because two $j\omega$ factors in the denominator). With relatively large values of R,

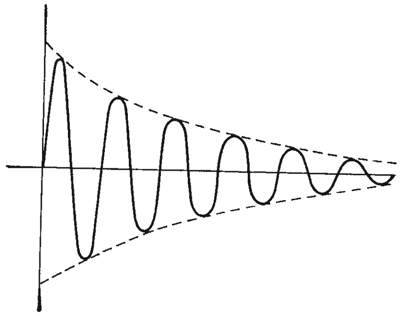


Fig. 6. Form of transient response revealed by Fig. 5.

both poles were on the $-\omega$ or imaginary-frequency axis, and we pointed out at the time that this corresponds to a non-oscillatory circuit; i.e., one having a unidirectional free response. As R was reduced, the poles moved closer to one another, eventually coalescing into a double pole (corresponding to " critical damping "). Further reduction of R made the poles separate vertically, as in Fig. 5; and we saw that the vertical distances, ascertainable from the time constants of the circuit, T_1 and T_2, represented to scale the frequency of natural oscillation of the circuit, denoted by ω_0. Being measured along the imaginary axis it is, according to our now well-established conventions, a real frequency. But because the poles also have a real-axis co-ordinate, $-1/2\ T_2$, representing an imaginary frequency, they may be said to represent complex frequency.

Poles on the axes, in fact, are special (though commonly occurring) cases; in general, poles and zeros can lie anywhere on what is called the complex frequency plane; i.e., the plane forming the groundwork of a pole-zero diagram. This idea of complex frequency is valuable both in theory and practice.

While we are at it, let us recall the form of the free response of a circuit represented by Fig. 5. It is a damped wave train, as in Fig. 6. I have dotted in the outline or " envelope," which is easily recognizable as the die-away or real-coefficient or (as we have just seen) imaginary-frequency exponential form. The waveform as a whole is obtainable by multiplying a continuous sinusoidal waveform by the die-away one, thus:

$$e^{\pm j\omega_0 t} \times e^{-t/2T_2}$$

(if we choose the coefficients of *t* from Fig. 5, according to our new-found principle that these represent frequencies, real or imaginary). According to the law of indices this can be re-written

$$e^{(\pm j\omega_0 - 1/2T_0)t} \quad .. \quad .. \quad .. \quad (7)$$

where the factor in brackets is the complex frequency, which reveals the rate of damping as well as 2π times the number of actual cycles per second, $=\omega_0$, in Fig. 6.

We saw, too, that if R was reduced to zero (keeping $1/T_1T_2$ or $1/LC$ constant) the poles continued to move in a semicircle until they reached the $j\omega$ axis. The frequency of the free response is then wholly real, so that an oscillation once started would go on indefinitely at the same amplitude. In other words, it is indistinguishable from that due to a sinusoidal source in a resistive circuit. That is why we changed the description " transient " to " free." Its frequency is of course equal to that represented by the diagonals in Fig. 5, joining the zero to the poles. Working this out by the application of Pythagoras to the co-ordinates of the poles, we get $1/\sqrt{T_1 T_2}$ as the answer, and from the definition of T_1 and T_2 this is equal to $1/\sqrt{LC}$, well known as the radian frequency of resonance of a tuned circuit to a source, which is equal to the frequency of natural oscillation

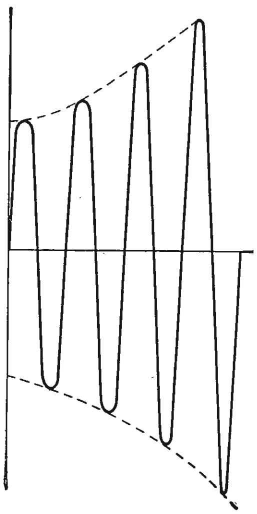


Fig. 7. If the poles in Fig. 5 had been to the right of the $j\omega$ axis, the response would have been like this.

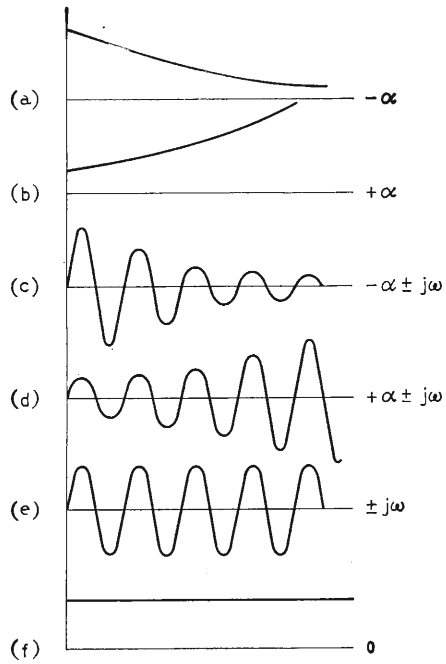


Fig. 8. Typical exponental curves corresponding to indices of the forms shown. Ordinary a.c. and d.c. are seen to be special cases.

when the damping resistance is reduced to zero.

Two months ago we stopped the poles here, but there is no reason why they shouldn't invade the area to the right of the $j\omega$ axis, completing a whole circle. This, obviously, must correspond to negative resistance, so when we plot a typical waveform, the counterpart of Fig. 6, we are not surprised to find that it is an exponentially growing one—Fig. 7. The real part of its exponential coefficient of *t*, unlike that in (7), is positive.

Just to consolidate these findings before we go on to re-consider forced responses, Fig. 8 shows the various free forms we have encountered, with the general mathematical forms of their complex frequencies. These, let us remind ourselves once again, are the constants by which *t* must be multiplied to form the index of e in the exponential functions from which the graphs in Fig. 8 are plotted, and they are also the co-ordinates of the poles and zeros (e.g., in Fig. 4 $-\alpha$ is $-1/T$). And again, imaginary co-ordinates signify real frequency, and vice versa. I have added the form when the index is zero, from which you observe that even d.c. is exponential. It is in fact intermediate between (a) and (b) and an extreme case of (e)—zero frequency.

We have now completely released circuit poles and zeros from confinement to any one axis; they can be *anywhere* on the complex frequency plane. And we have closely related them to the form and frequency of free responses in the circuits to which they relate. What about sources?

We began by considering only sinusoidal excitations, which accounts for the fact that points representing their frequency were confined to the $j\omega$ axis—e.g., the dot in Fig. 4. But excitation could be made to take any of the forms shown in Fig. 8; not just (e) or (f). Their frequencies would be complex, and dots representing them could come anywhere on the complex frequency plane.

If we chose form (a), then the frequency dot would lie on the $-\omega$ axis. By choosing a suitable rate of die-away, we could make it actually coincide with the pole there. In other words, we could make the distance between the two dwindle to nothing. But, as we saw right from the introduction of poles, this distance represents the denominator of a circuit function—in this case its admittance. So a zero value means that the admittance is infinitely large. To anyone who thinks in terms of sinusoidal excitation, this clearly implies that *any* voltage, however small, will cause an infinitely large current. But can we really see a voltage of the form shown in Fig. 8(a) causing an infinitely large current in a heavily resistive circuit of the Fig. 1 type? Now if the pole had moved up to the dot on the $j\omega$ axis, representing a sinusoidal excitation, that would have meant zero resistance and an unlimited growth of current as in Fig. 8(d), which is just what one would expect. But the reverse kind of movement, dot to pole, seems to have landed the pole-and-zero idea in an absurdly impossible situation. The hero, so to speak, is villainously bound hand and foot to the loaded inter-continental ballistic missile now rising from the launching pad, and how can even the most resourceful author possibly rescue him? The time has obviously come for reminding you to make sure of receiving next month's *Wireless World.*

LETTERS TO THE EDITOR

The Editor does not necessarily endorse opinions expressed by his correspondents

Television Interference

THE proposed new TV stations for increasing B.B.C. coverage fill me with misgivings.

Is enough consideration being given to the problem of TV receiver oscillator radiation and its effects?

With the present-day general use of vision i.f.s around 35Mc/s, any set tuned to Channel 4 causes 2nd-harmonic oscillator interference to be superimposed on nearby Channel 9 receivers, particularly in fringe areas. Similarly with Channel 5 and Channel 11.

I note that the problem is becoming imminent in Jersey (Channel 4 and Channel 9).

One wonders just how many more cases there will be when the new Stage 3 is complete!

Peacehaven, Sussex. R. G. YOUNG.

N.B.

(a) $2(61.75+35)=193.5$ Channel $9=194.75$.
(b) $2(66.75+35)=203.5$ Channel $11=204.75$.

[Frequencies can hardly be shifted to avoid oscillator radiation. Receiver radiation must be reduced.—ED.]

V.H.F. Aerial Systems

I WOULD like to comment on finding the antinodes for all three B.B.C. v.h.f. services, as commented on by "Free Grid" in the April and June issues.

Most radio dealers know of this, but the finding of the position takes more time than the client is generally prepared to pay for. The exception is where the dealer has a field strength meter, though the finding of the optimum position so often is useless because the antenna cannot be fixed there for one reason or another. So a less useful position and a directional array are chosen.

This appears to be under fire, but believe me it is good practice to use directional aerials for v.h.f., f.m. One of the snags of domestic receivers is that they are made to a price limit, which results in poorer discriminators than we would wish for. With the concentration of aircraft over this country nowadays, anything that will narrow the area in which an aircraft needs to be to cause the reflected signals that give rise to flutter, and multipath distortion is to be recommended.

While I am sure all the above is well known to you, I feel that the publishing of incomplete articles may lead the unwary into difficulties, and I feel, Mr. Editor, that Mr. Free Grid was, in fact, right when he said that opinion believes the service has not quite come up to that claimed for it. Mind you, there were some who didn't expect it to! Nevertheless, I should hate to be without.

Coulsdon, Surrey. W. T. CLEGG.

Meters and Senses

REFERRING further to the letters from "Cathode Ray" and Mr. Leslie Hills (May and June issues), the subjective effect of harmonically complex waves varies to such an extent with different observers that no exact evaluation has ever been possible. Removing the fundamental from any sound having properly related harmonics results, as "Cathode Ray" says, in synthesis by the ear of the pitch note. But this is only true if the harmonics are in tune. In the simple example of a small upright piano at a little distance, it is quite impossible for the extreme bass strings to radiate true low pitches because they just can't excite enough air. Yet one can hear the bass. However, if the piano is *not* in tune, one cannot hear the bass but only a discordant sound and the ear is irritated and frustrated.

It is well known that complex tones are only representative in a general way. There is no such thing as an exact oboe tone—only a kind of sound spectrum which is sometimes dependent almost entirely on the context of other notes for its true classification as a tone value. This is true of every complex-tone instrument, because only over a small range of pitch and power is the true designed sound engendered.

A great difficulty is encountered by the amateur in tuning musical instruments, where it is essential that the fundamental be correct, if a complex tone source is used. It is sometimes quite difficult to tell which octave one is tuning. This happens if the 2nd and 4th harmonics are very powerful, as they are in many reed or "brass" instruments. For this reason, a pipe organ tuner will always tune against a simple, loud tone —generally a 4ft principal. Beats are then easily heard.

Mr. Briggs must surely know that the effect of ambient temperature changes on the pitch of musical instruments is very marked, according to their material of construction. This is why the oboe is used as an orchestral tuning standard, since the cane reed and wooden tube are almost unaffected by temperature changes; but they are greatly affected by humidity changes, whereas the brass is not; so that we find a constant rising and falling of pitch in a random way between different sections of an orchestra during a performance. Observations made by Philips* show a variation of 12c/s either side of A=440c/s during a performance.

When the ear anticipates a sound it has not yet heard, it is very critical of pitch; but, after the sound has existed for several minutes, its resolving power diminishes and this does not seem so important.

Any alteration in the relative humidity of the atmosphere in a concert hall will affect the attenuation of sound, this becoming greater as the frequency increases. The presence of CO_2 also affects the velocity of propagation, and some or all of these effects are usually present in a concert hall. We do not notice them because the hearing system is so wonderfully self-adjusting.

Any real variation in the relative "smoothness" or "harshness" of sounds must be due to either the hall acoustics, the loudness levels, the number of instruments used, or the conductor's technique of expression or phrasing.

Nottingham. ALAN DOUGLAS.

* *Wireless World*, 11 May 1939, p. 441 and *Philips Tech. Rev.*, Vol. 4, No. 7, July 1939, pp. 205-210.

Transistor Circuit Conventions

NOTWITHSTANDING recent judgments based on statistical trends amongst your correspondents, I would venture a less politic but more traditional compromise based upon a restatement of axioms.

1. Transistors are essentially current-operated devices; so that "the current's the thing" if we want to understand how the circuit works. This allows the use of arrows in the connections and can eliminate the need for pluses and minuses associated with voltage levels.

2. For consistency the current should go principally in one direction in any one diagram—this is analogous to the same requirement in respect of signal flow. The flow of current in the devices can be thought of as the "how" of the circuit and the flow of signals could be described as the "why" of the circuit—the tactical and stategic, circuit and system, engineer and project leader, repair and maintenance aspects of our vocation.

Since signal flow usually goes across the page the current flow should be vertical, so enabling us to get two lots of information on to one diagram.

Unlike the valve which "violates" Kirchhoff's Law the transistor does have currents in every arm. And unlike valves again, this current can be in either direction, both because of the way the circuit may operate and because of complementary devices.

A unique current direction can be specified, of course, and this is in the base connection.

Drawing conclusions from the foregoing statements and applying them to the concept of a circuit diagram that is to have optimum heuristic values, we get:—

(a) A master arrow pointing vertically on the diagram shows which way the current flows in the circuit.

(b) The transistor symbol has an arrow in the base lead only, whatever symbolism is evoked for its representation, pointing in for p-n-p and out for n-p-n.

Comment may be in order concerning waveforms (unfortunately restricted in the nature of things, still, to voltages). My colleagues remind me that in the early days my bench was distinguishable by the scope sitting upside-down whilst I grappled with "hole" currents, etc., produced by available transistors. In these enlightened days we are beginning to free ourselves from the trammels of signals measured exclusively from something called "earth" or even "common", and differential inputs to our scope amplifiers enable us to get waveforms of either polarity.

Finally on the subject of "emitters" and "collectors". These terms apply in certain traditional configurations and will continue to distinguish usefully various parts of the attached circuit. But there need be no basic distinction in either manufacture or use, and the pivotal significance of this historically useful concept should accordingly be reduced or eliminated. The point might be made by considering the gain in reliability and repeatability that would accrue to a circuit which had been so designed that it would function adequately with its emitters and collectors interchanged.

Specifically I suggest that polarity is no longer a major preoccupation and as such can be chosen, i.e., left free, to complement some other feature of the design; that a transistor symbol be adopted showing principally base current; and that all diagrams carry an embryo legend containing an explanatory list of the conventions adopted by the designer and/or drawing office.

Cowes, I.W. R. MATTHEWS, Decca Radar Ltd.

THE circuit-drawing conventions discussed since your January issue seem to me, with one bright exception (Messrs. Knowles and Braithwaite in the April issue), to disregard wholly the less bright brains constituting the undisputable majority of schematic *users*—the technicians and service men. Highly trained men are unquestionably able to *read*—literally—a schematic, no matter how it is drawn. Nevertheless, even the theory-inclined of your correspondents admit that they are in the habit of redrawing a circuit for closer analysis. This discloses the need for universally accepted circuit-drawing conventions, which naturally (to me at least) must fill the following demands: (a) easy to grasp circuit *function*. (b) theoretical analysis possible without redrawing and (c) easy to draw.

The first point emphasizes the *user*, who by "pattern recognition" will be able to do his job faster and with more confidence and less stress. Of course, new patterns will be developed as time passes, but if presented in much the same way as earlier schematics, function will be more easily interpreted. This is the point of greatest importance and here is where polarities come in, both the diagram as a whole and in the waveform inserts. To me it seems clear that no special importance is to be placed on whether + or − is uppermost in the diagram—the pattern recognition is of far greater importance. The present trend not to join the d.c. source terminals with the appropriate points of a stage takes away a great deal of unnecessary and in no way enlightening lines and crossovers in our diagrams, *provided the arrangement of the d.c. source is clearly defined in the diagram.* If so, even scores of arrowheads (with polarity symbols) cannot confuse but, better still, help in analysing the function. On the other hand, the waveforms ought to be presented with positive upwards, once and for all—shifting the wires to the scope is easier!

As for point (b), one must take account not only of the designer's needs, but also, and more important, of the teaching demands. Realizing that the average designer most probably is going to use the schematic-drawing philosophy absorbed during his training years for the rest of his life, the importance of this point is clear. Since a good deal of teaching effort is aimed at theoretical treatment I cannot see why there should be any controversy hidden in demanding uniform principles for schematics.

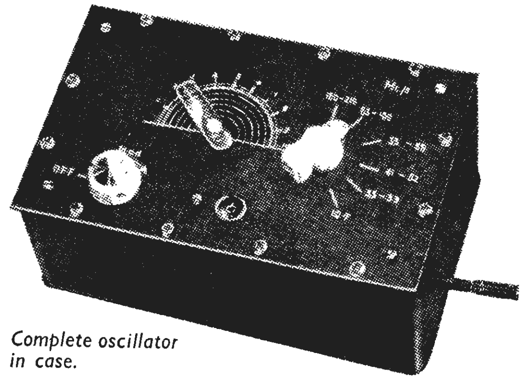
The last point may seem to be of less importance, but is it so in real life? The many technicians and servicemen who use diagrams are now faced not only with the tastes of the draftsmen at every factory, whose devices they have to handle, but also the highly individually drawn schematics published in journals and which the man in the field on behalf of self-education is more or less compelled to get acquainted with. Clearly the individual as well as the community as a whole would be able to benefit by a unified convention in schematic drawing, and naturally the least trouble-making one that fulfils the needs in other respects.

In conclusion I am in favour of the "collector up"-convention in conformity with the reasons indicated above, and moreover I hope to see a convention growing out within the next quarter-century!

Stockholm. O. HEDSTRÖM.

TRANSISTOR DESIGN FOR HIGHER FREQUENCIES

By Sqn. Ldr. D. R. BOWMAN



Complete oscillator in case.

PORTABLE TEST OSCILLATOR

WHEN checking receivers and other electronic apparatus " on site " it is often inconvenient to have to use mains-attached equipment. The professional serviceman obviously needs to be properly equipped; but where the only transport available is a private motor car, one hesitates to put one's laboratory apparatus in the boot and the possession of alternative highly-portable instruments has definite advantages. A brief survey over the past year of calls by friends for electronic first aid showed that the signal generator was by far the most-used item. and it was therefore decided to attempt the development of a transistorised test oscillator of the " pocket " class, covering all the frequencies of interest with v.h.f./f.m. and television.

The instrument described is not intended to be a substitute for a proper laboratory signal-generator. Considerations of small size and portability rule out any serious attempt to provide highly accurate scales and it is hardly feasible to incorporate a meter for monitoring power output. The type of attenuator used is capable of high calibration accuracy; given " precision " workshop facilities it is possible to construct a sub-standard component of this type. This is not attempted here: if accurate measurements need to be made—as in development work—an instrument designed for the purpose has to be used.

Nevertheless, within its limitations, the portable test oscillator has features to commend it. Its resetting accuracy is sufficient to enable a receiver to be aligned fully from random starting conditions. It produces a stray field that is negligible at all frequencies and the very low minimum-power output enables the sensitivity of a receiver to be estimated. Because only small variations of power occur within any one range, as the tuning capacitor is adjusted, it is possible to plot a receiver response curve with some confidence—although, for reasons to be given later, it has not been found possible within the present design to add a frequency-sweep facility.

Because a fairly open scale would be required if even approximate calibration was to be attempted, the use of a tuning capacitor of small maximum capacitance was thought to be best; warming-up drift is not a problem with transistors, and it was desired to reduce to a minimum circuit capacitances. Excessive circuit capacitances would restrict the tuning range on any band of frequencies. Provision was made to cover the frequencies shown in Table 1.

TABLE 1

Range	Frequency (Mc/s)
1	10.2— 11.2
2	33 — 39
3	41 — 52
4	52 — 68
5	85 —100
6	180 —210

The desired switching could then, as will be seen later, be effected with a single-wafer double-pole switch.

A review of available transistors showed that several suitable types could readily be procured, and the remaining components—even when only of the smaller standard sizes—would enable a very light and compact instrument to be devised. The attenuator presented a difficulty; but the solution to the problem was found in a piston-type attenuator requiring only moderate skill to construct with sufficient accuracy.

R.f. Oscillator

It was considered essential to simplify switching so that only one two-pole wafer would be needed: consequently the negative-resistance oscillator was investigated. This type of oscillator would also have the advantage that if, in the future, it became necessary to extend the frequency range of the instrument into Bands IV and V, the modification would consist only of replacing the Band-I inductors by simple transmission lines.

Fig. 1(a) shows the simplified diagram of a type of negative-resistance oscillator, which will be seen to be of the grounded-base configuration. Fig. 1(b) shows the equivalent circuit at high frequencies. In writing the equations for this circuit some simpli-

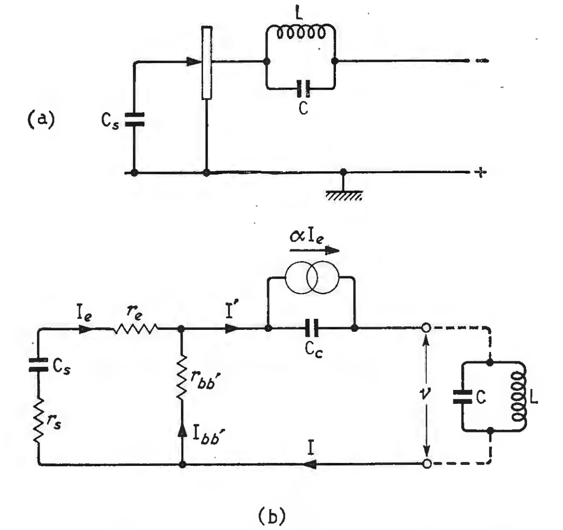


Fig. 1 (a) Simplified, grounded-base negative-resistance-oscillator and (b) its equivalent circuit at high frequencies.

fication can be effected by neglecting emitter resistance and source series resistance.

$$I = I_e + I_b$$
$$= I_e [1 + (jX_{Cs}/r_{bb}')] \quad .. \quad .. \quad (1)$$
$$I' = I - \alpha I_e$$
$$= I_e [1 - \alpha + (j\, X_{Cs}/r_{bb}')] \quad .. \quad (2)$$
$$v = v_{rbb}' - v_{Cc} \quad .. \quad .. \quad .. \quad (3)$$
$$Z_{out} = v/I \quad .. \quad .. \quad .. \quad .. \quad (4)$$

The complete analysis is tedious[1], but it turns out that when the input impedance is capacitive (X_{Cs} negative) the output resistance is negative provided X_{Cs} is greater than $(\omega/\omega_c)r_{bb}'$. Thus oscillations will be set up in a high-Q inductive load. Obviously also it is not practicable to reduce C_s to too low a value, otherwise the source impedance is virtually only r_{bb}' (input open-circuited) and $Z_{out} = r_{bb}' - j/\omega C_c$ where no condition exists for the output resistance to become negative.

At cut-off frequency therefore $X_{Cs} = r_{bb}'$ and putting in a few practical values, such as $r_{bb}' = 75\,\Omega$ and f = 100Mc/s, C becomes about 22pF as a maximum value. Above cut-off frequency the required capacitance diminishes to say 10pF at 200Mc/s. In any practical circuit some of this would be contributed by circuit strays and, of course, at higher frequencies the inductance of the base connection affects the circuit conditions. Capacitances for lower frequencies would be expected to be in the same proportion; thus at 10.7Mc/s something under 200pF would be expected. In practice it was found that these considerations gave a very reasonable guide to the actual values needed, which were by no means critical in any event. Suitable transistors are Texas Instruments 2G101 and Mullard OC171 diffused types.

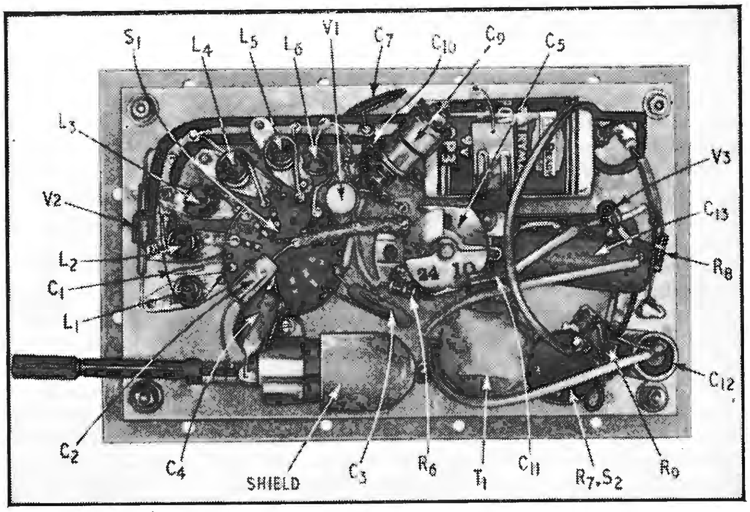
A.f. Oscillator

The intended power supply is an important factor to be considered in deciding the type of oscillator to employ for the modulating stage. Here a miniature 9-V battery (Ever-Ready type PP3 or similar) was envisaged to supply both r.f. and a.f. oscillators, and economy of current consumption indicated a single-transistor stage if possible. In a single-stage R-C oscillator of the three or four-section type the effective current gain of the transistor must be high enough to overcome the attenuation of the phase-shift network, even with a partially discharged battery. The input impedance in the grounded emitter configuration is low, and serious mismatch to the network is inevitable; consequently not many transistors are suitable for this kind of circuit[2]. A simple feedback oscillator, using a Mullard type OC75, of the tuned-collector type was therefore developed. The waveform was found to be reasonable developed. The waveform was found to be reasonable[3].

Pure amplitude modulation of the r.f. oscillator can hardly be arranged in any simple fashion, but it was found that injection of the modulating current at the base of the r.f. transistor gave reasonable freedom from frequency modulation. A high resistance connecting the collector of the a.f. oscillator and the base of the r.f. oscillator ensures that the a.f. oscillator acts as a constant-current generator, besides reducing the audio current to an appropriate figure.

Attenuator

The attenuator used has to be effective from 10.7Mc/s to 200Mc/s—with possibly an eye to the future when it might be needed to work well enough at over 600Mc/s. At higher frequencies it is difficult to ensure that any resistive variable attenuator would



Interior of prototype oscillator. Main components are indicated and, in general, signal leads must be kept short and be of heavy-gauge wire (vide S_{1A} to V_1 collector and tuning capacitor). (Slight differences between photograph and drawing (Fig. 5) should be ignored.)

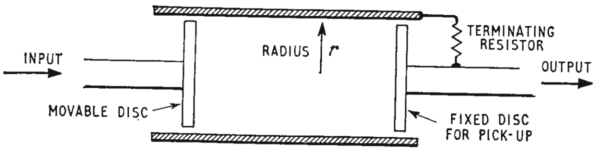


Fig. 2. Principle of piston attenuator.

be effective. The piston attenuator is the obvious alternative, and has the advantage of giving a relatively high attenuation in a small physical space. Where a large maximum power output is needed its big disadvantage is that its insertion loss is very high: however, a large output is not required for the intended application, and preliminary rough estimates showed that a useful signal would be obtained. In practice it was found that in its minimum attenuation position enough power was available to saturate even a roughly-aligned receiver, even giving recognisable signals through a single detector stage followed by the usual type of audio amplifier. Thus stage-by-stage alignment *ab initio* need present no problems.

The piston attenuator is, in effect, a waveguide operated at a frequency below the cut-off frequency. Two types exist; in the mutual-inductive type a wave is launched into a waveguide by means of a coil, pickup being effected by a second coil connected to the output terminals, while in the mutual capacitive type the excitation of the waveguide, and pickup, are effected by capacitive probes—usually disks.[4] The capacitive type (Fig. 2) is used in this instrument. The electrode launches a TM_{01} mode in the guide, and the attenuation for this mode is given by the formula:

$$\alpha = (2\pi/\lambda_c)\sqrt{[1-(\lambda_c/\lambda)^2]}$$

where α is the attenuation in nepers/metre and λ_c is the waveguide cut-off wavelength.

For a circular waveguide of internal radius r this reduces to $20.9/r$ dB/m when the ratio λ_c/λ is much less than unity, r being measured in metres. It is common practice to terminate the output disc with a resistor of value equal to the characteristic impedance of the cable to be used—almost certainly 75 to 80Ω in practice—and is recommended here because the total length of cable used is very likely to be comparable with a quarter wavelength in linear dimensions.

The waveguide chosen in this instrument is a brass tube of internal diameter $\frac{9}{16}$in and about $1\frac{1}{4}$in in length. This is of small size compared with the shortest wavelength generated and an almost constant attenuation factor is obtained over the frequency ranges covered. The attenuation expected is about 30dB/cm movement of the pickup disk, and in all about 90 or 100dB. In practice this expectation was achieved.[5]

Stabilization against Battery-voltage Changes

For calibration to remain usable when the power supply voltage changes it is necessary to stabilize the source of direct current. Suitable Zener diodes are readily available: the author used S.T.C.'s type Z2A68 $\pm$ 20%. Although the battery may be used when discharged to as low as 5V the Zener diode specified is then out of action. Thus replacement is recommended as soon as calibration begins to be lost; a very useful life will still be obtained from the battery.

F.m. Facility

In case it was required to use the test oscillator with an oscilloscope for the display of amplifier response curves it would have been well worth while to complicate the assembly slightly by the inclusion of a reactance diode in the tuned circuit. Experiments carried out on these lines gave disappointing results. A variable-capacitance diode, type SVC 1, was available for tests, and its measured capacitance with variable voltage—shown in Fig. 3—appeared to be suitable for the application. It was found that while reliable oscillation could be maintained in the tuned circuit over part of the range of capacitance, at the lower voltage end of the sweep the tuned circuit was apparently loaded so heavily that oscillations stopped. While this could have been avoided by limiting the sweep, the provision of all the frequency

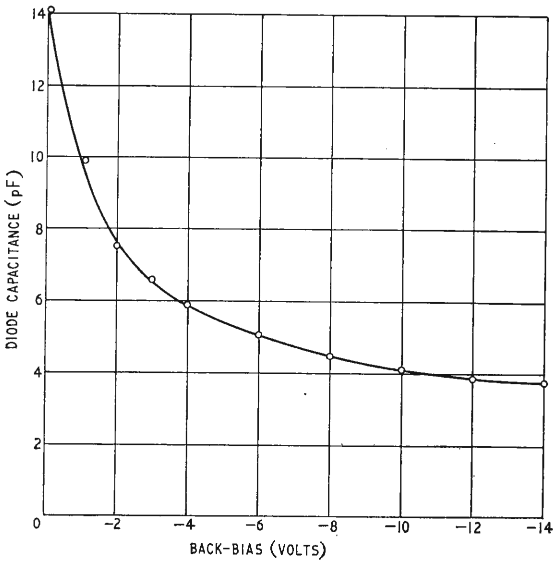


Fig. 3. Back-bias/capacitance characteristic for type SCVI diode. Measurements were made at 10Mc/s.

ranges needed called for about ten ranges, necessitating two wafers for the switching.

Construction

The circuit diagram is shown in Fig. 4. It will be noted that a single, adjustable emitter capacitor is provided for the Band-II and Band-III ranges; a capacitance can readily be found which gives good results on both these ranges. Fixed capacitors were originally tried; but transistor " spread " and manufacturing tolerances in capacitances gave rather inconsistent results. It is considered that circuit " strays " will probably be not reproducible enough for account to be taken of them accurately. Fixed capacitors switched into the emitter circuit were found to be non-critical for the lower frequency ranges, as might be expected. Capacitors of 120pF and under are silver-mica types: ceramic dielectric types are suitable for the larger sizes with paper for those over 0.10μF. The tuning capacitor is Jackson's type C804 and the trimmer is a Philips "beehive" of 2 to 10pF.

Printed-circuit construction is employed for ease in working and because the copperclad laminate (Bakelite Grade E60) can be silverplated readily. Only the minimum area of copper is etched away so that an effective "chassis" of low r.f. impedance is provided to which soldered connections can readily be made. Fig. 5 shows the etch pattern recommended. There are few conductors and spacing is not a problem, so it will be found simple enough to paint the conducting surfaces out direct on to the laminate, using a fairly thin shellac or cellulose varnish as the acid resist. 30% ferric chloride solution (by weight $FeCl_3$) with 1% concentrated hydrochloric acid added, gives a quick etch, taking about six to ten minutes at room temperatures with moderate agitation of the etch bath. If simple apparatus is available the circuit board may be silver-plated with advantage. If not, it may be varnished with shellac after all soldering has been carried out, but the amount of space for the manipulations involved is not great.

The two components to be made are the transformer in the audio circuit and the attenuator. These will be discussed in turn.

A.f. Transformer

There seemed to be nothing readily available commercially to serve in the a.f. stage and a small component was wound using oddments. A few Silcor I-laminations may be cut up to make a small core $\frac{3}{4} \times \frac{3}{16}$ in square, one single lamination being double length and bent over at right angles for later mounting in an upright position. A bobbin is made from thin card to fit this core, and the following windings are placed on it:

800 turns 40 S.W.G. enamelled-copper wire, tapped 300 turns from one end.
A single layer of tape.
40 turns 24 S.W.G. enamelled copper wire.

The whole is then dipped in shellac varnish to obtain rigidity and to exclude moisture, and then allowed to dry thoroughly; this takes about 48 hours in a warm place. The windings are "scramble-wound" as voltages are low and insulation is not a difficulty.

Attenuator

The attenuator requires more care in construction, and is best assembled direct on to the circuit board and aligned by trial and error. The essential dimensions are given in Fig. 6.

The waveguide part of the attenuator consists of a brass tube $1\frac{1}{4} \times \frac{3}{16}$in internal diameter. This is first carefully cleaned inside and out with sandpaper, and is silver-plated internally. Two brass disks are obtained, each less in diameter than the tube, and are drilled and tapped centrally to accept a 4-BA brass screw. The brass screw is inserted and secured in place with solder, which should not be excessive in amount. Each disk is then coated thinly with varnish and is allowed to dry thoroughly. Shellac may be used (two or three coats), but in the prototype contact adhesive was used in a single coat.

The position of the coaxial output socket is marked out and holes drilled. When the socket has been fixed in position the waveguide is arranged in a horizontal position by means of a circular clip (from an electrolytic capacitor) fitted with a mounting lug. Its height above the circuit board may be adjusted, using washers as necessary, so that the centre line lies a very little above the end of the coaxial connector inner. The coaxial socket is then loosened so that one disk, with its screw attached, can be inserted in the end of the tube. A thin strip of bakelised paper or acetate sheet is then introduced between the edge of the disk and the inside of the waveguide tube; this is forced into position to hold the disk very tightly. This step is necessary so that when a coaxial plug is wriggled into the socket—as is usual—it causes no displacement of the disk. The socket is then tightened up. If adjustment has been carried out accurately enough the brass screw will now just touch the socket inner conductor; the two are then soldered together with plenty of solder to form a strong mechanical joint. This operation should be

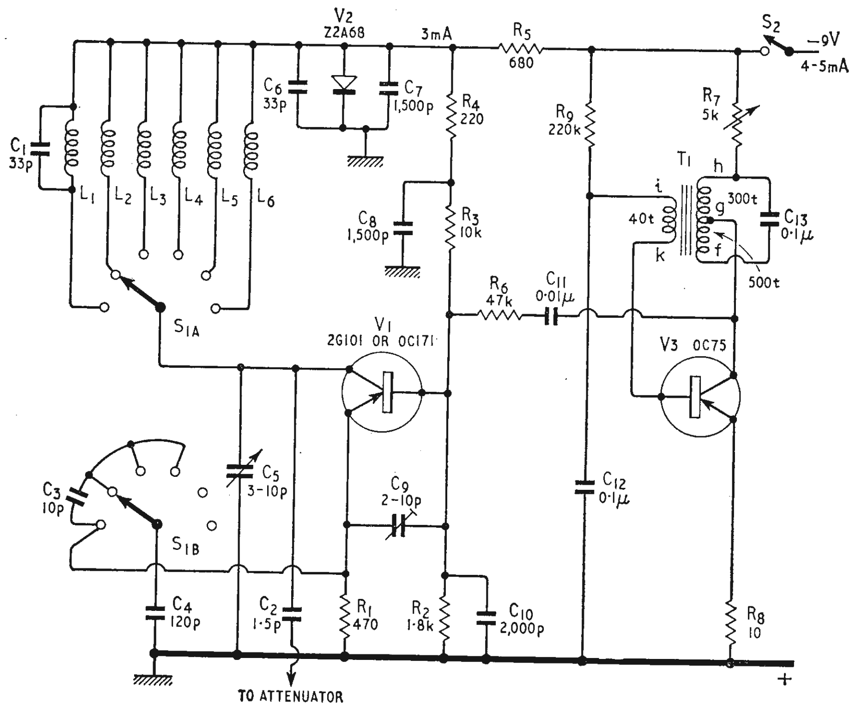


Fig. 4. Circuit of portable test oscillator.

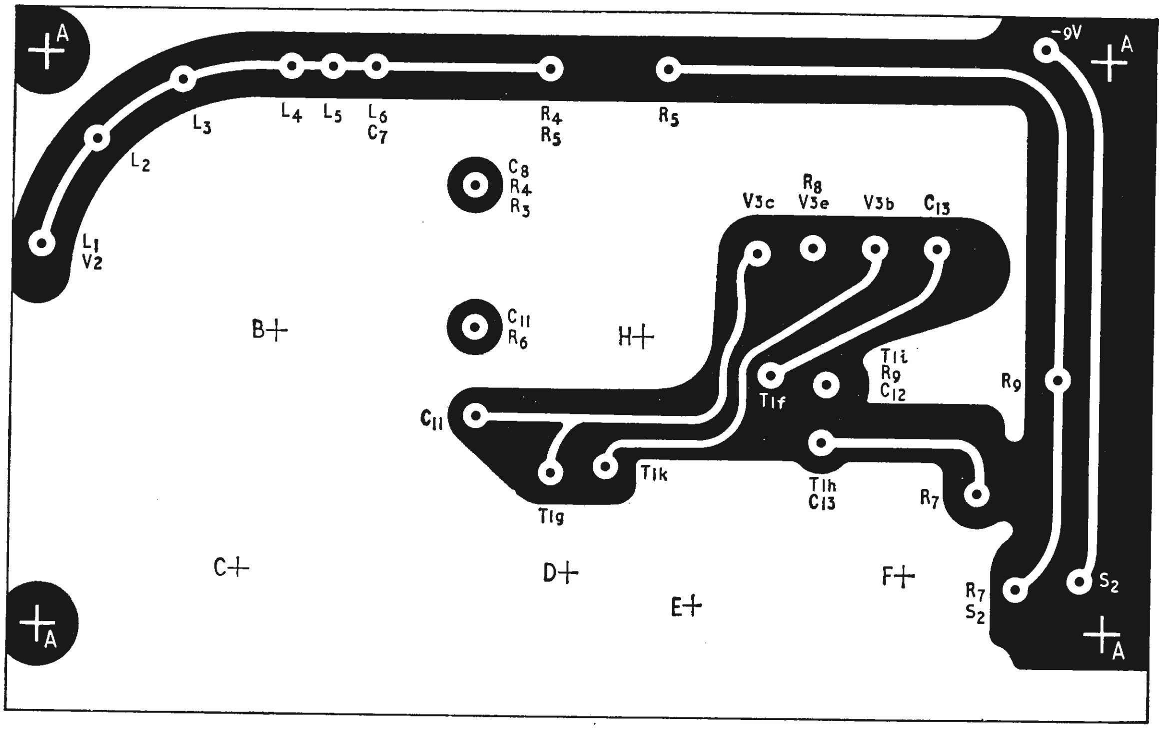


Fig. 5. Printed-wiring layout, full size for final version of oscillator. Mounting centres are marked +:—A, screws to case; B, S_1; C, stand-off insulator for movable attenuator disk; D, coaxial socket; E, T_1 mounting; F, R_7 and S_2. "Earthy" connections are taken to nearest point on board.

carried out with a very hot iron, preferably using thermal shunts to avoid overheating the insulation of the socket or of the disk. At the same time a 75 or 80-Ω resistor—of small physical size to minimize self-capacitance—is connected between the coaxial socket and waveguide, the point of attachment to the tube being at the edge of the tube itself.

The other disk is now given a second coat of contact adhesive, especially round the edge so as to build up a thin "tyre." When dry it should be tested for a very loose fit inside the tube; for this reason it would be well to start with an undersized disk, about $\frac{1}{16}$in less in diameter than the tube.

A long brass nut, or two thinner ones with a lock-washer between, should now be soldered to the end of a ceramic stand-off insulator about $\frac{1}{2}$in in length. This assembly is now so located on the circuit board that the centre line of the nuts lies on the centre line of the tube and about $\frac{1}{2}$in from its end. A suitable hole for fixing this component will be one a little oversize so as to allow for lateral adjustment as needed. The brass screw is now screwed into the nut and the disk inserted into the end of the waveguide tube. The stand-off insulator is now fixed in position. If all has gone well the screwed brass rod should rotate freely, carrying the disk easily in and out of the tube. Adjustment for height may be made with washers as necessary. The procedure may seem a little complicated but good results can be had with moderate skill if due care is taken.

A piece of Paxolin tube is next fixed to the free end of the screw to afford an insulating handle for adjustment; it may be made secure by pinning, or by wedging aided by adhesive. Alternatively an ebonite rod may be drilled and tapped to screw over the brass rod.

A check should now be made to see that the two disks can approach each other as closely as their insulation layer permits. Finally a screen is fashioned out of bent-up tinplate just big enough to cover the coaxial inner conductor and the end of the waveguide. This is now soldered to the circuit board at several points.

Calibration

On all but the 10.7-Mc/s and 33-39-Mc/s ranges the calibration is a simple matter because a group of highly accurate frequencies is available for each

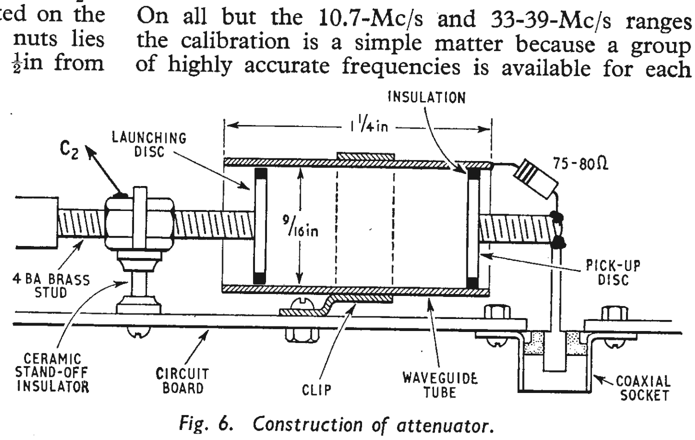


Fig. 6. Construction of attenuator.

range in the shape of transmitted programmes. Using a receiver as indicator, points may be plotted and a graph drawn which gives reliable markings on the scales used. For the 10.7-Mc/s and 33-39-Mc/s ranges an external calibrated signal-generator will be required, and the indicator may still be a suitable receiver, using the i.f. amplifier only. The use of the television receiver screen as a visual indicator is not difficult in practice, as will be found with a little experiment. It is recommended that the 10.7-Mc/s scale be placed nearest to the spindle of the tuning capacitor, the ranges working outwards. In this way the scales requiring greatest accuracy of marking will be the most extended. It is doubtful whether hand-marking of the scales can be carried out to better than 5%, even with care; but this is unlikely to be an important limitation on the use of the instrument for its intended purpose.

Performance

The power output obtained differs as between ranges but the difference is not embarrassing. However, the temptation has to be resisted to adjust the emitter capacitor for maximum power on the highest-frequency ranges: there is a range of values with which reliable oscillation can be obtained and the aim should be to use the middle portion of this range. When so adjusted the power into 80Ω was as shown in Table 2.

TABLE 2

Frequency (Mc/s)	Power (μW)
10.7	0.5
33 — 39	1.0
41 — 52	1.1
53 — 68	1.4
85 —100	2.0
180 —210	2.8

The amplitude varied a little over each range, to the extent of about 10% on the lowest two to a little over 20% on the highest range. If very thorough internal shielding of the attenuator input were arranged, together with perhaps a resistive coupling between the tuned circuit and the attenuator input, this variation might be reduced. However, the input capacitance of the attenuator is not small and a few experiments indicated that a resistive coupling caused such loading of the tuned circuit that oscillation was sometimes not reliable. For the intended application of the instrument it was thought that the variations of power might be accepted.

Shielding

The degree of escape of radiation and induction field must be very small if an oscillator is to be useful. The printed circuit is therefore mounted inside a metal case whose lid consists of a small panel of aluminium about 16 or 18 s.w.g. In the prototype a suitable tin with soldered seams was selected, the edges turned over and the aluminium panel attached with twelve self-tapping screws. The circuit board is insulated from the case except at a point close to the coaxial socket, where a connexion is made by means of a stout piece of braid. Circulating currents are thus avoided, and the only appreciable leakage is by way of the coaxial inner and, to a smaller extent, the holes carrying the control spindles. The stray field is insignificant compared with the intended output, and if the attenuator is in a position near its minimum stray field can hardly be detected.

Coil Details

All coils except L_6 are wound on $\frac{9}{32}$-in diameter polystyrene formers, iron-dust or brass cores are used as needed.

Inductor	Frequency (Mc/s)	Winding Details
L1	10.7	32 turns, closewound, 28 s.w.g. enamelled copper wire (en.).
L2	33— 39	14 turns, spaced by wire diameter, 28 s.w.g. en.
L3	41 — 52	12 turns, spaced by wire diameter 28 s.w.g. en.
L4	53 — 68	9 turns, spaced by wire diameter, 28 s.w.g. en.
L5	85 —100	6 turns, spaced to occupy 0.4in., 18 s.w.g. bare copper (silver-plated if possible.)
L6	180 —102	$1\frac{1}{2}$ turns, adjacent part turns spaced to occupy 0.4in. 18 s.w.g. bare copper (silver-plated if possible).

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Commercial Literature

Miniature tungsten filament lamps: details of more than 1,400 types and quick-reference folder of classes of lamp available from Vitality Bulbs Ltd., Neville House, Neville Place, Wood Green, London, N.22.

Racking and cases for electronic equipment are described in 10 languages (English, French, German, Spanish, Italian, Portuguese, Dutch, Swedish, Norwegian, Danish) in Imhof's International Catalogue. Alfred Imhof Ltd., 112/116 New Oxford Street, London, W.C.1.

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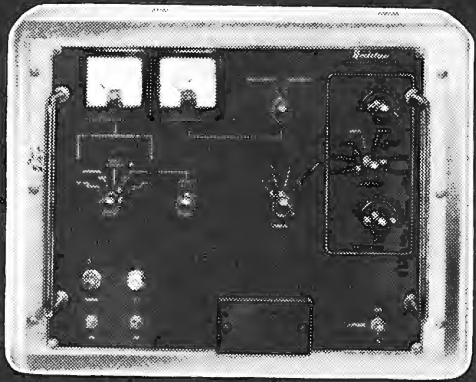

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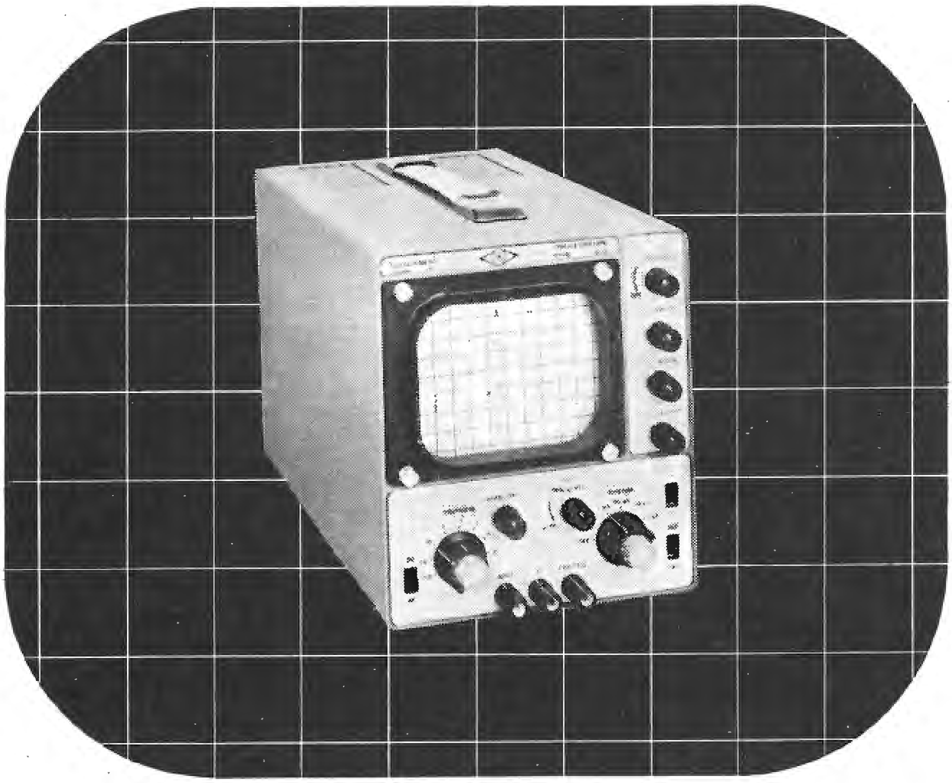

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The Boucherot Effect

By J. F. YOUNG, C.G.I.A., A.M.I.E.E., A.M.Brit.I.R.E.

IN most cases when one is confronted with a series tuned circuit such as that of Fig. 1, it is safe to assume that it is being used because of its frequency selective properties. However, this is not always so. When operated at its resonant frequency, the series tuned circuit has another important property, called the Boucherot effect, which has found many applications. Because this effect has been utilized mainly in the electrical power industry, the fact that it can occur in electronic circuits tends to be overlooked and it is rarely mentioned in electronic literature.

What is the Boucherot effect? To answer this we must ask another question. In Fig. 1 an inductor L is connected in series with a capacitor C to a sinusoidal supply of voltage V$\sin \omega t$, ω being 6.28 times the supply frequency. A load of impedance Z is connected across the capacitor. If the capacitor and the inductor resonate at the supply frequency, what is the current in the load Z?

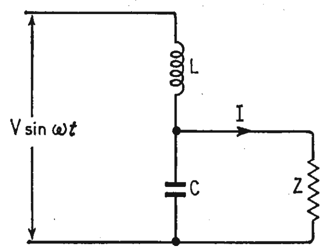


Fig. 1. Basic Boucherot constant current circuit.

In 1919, Boucherot had his own ways of solving this problem, since he did not believe in the use of the j operator to indicate 90 degrees vector rotation[1]. In fact he compared the j operator with a sausage machine—" one can see what goes in and what comes out, but what happens in between times is a mystery! " Whether, like Boucherot, we use differential equations to analyse his circuit or we use the j notation which he disliked, the value of the load current can be obtained very easily if we make use of Norton's theorem.

Constant Current Circuits

If the load is shorted out in the arrangement of Fig. 1, the current in the short circuit is simply:—

$$I = \frac{V}{j\omega L}$$

Now Norton's theorem states that, in effect, this short circuit current flows into a network consisting of the load in parallel with the internal impedance of the source, so that application of this theorem to the circuit of Fig. 1 gives the arrangement in Fig. 2.

The internal impedance of the Boucherot circuit is formed from the inductor L in parallel with the capacitor C, and since we have specified that these resonate at the supply frequency the internal impedance is infinite. This follows from:—

$$Z_{int} = \frac{j\omega L}{1 - \omega^2 LC}$$

which is infinite when $\omega^2 LC = 1$ at resonance. Since the inductor and the capacitor in parallel in Fig. 2 form an infinite impedance, the whole of the current I must flow through the load Z. By Norton's theorem, therefore, the current in the load Z in Fig. 1, as in Fig. 2, is equal to the short circuit current $I = V/j\omega L$. The surprising thing to notice here is that the load current is completely independent of the value of the load impedance Z.

Thus the circuit of Fig. 1, when supplied at its resonant frequency from a constant voltage source, gives a constant current in the load regardless of the nature or value of the load impedance. The load current depends only on the values of supply voltage, inductance and frequency, and it is phase shifted by 90 degrees to the supply voltage. In effect, the circuit acts as a constant current generator or, to be more exact, as a constant voltage to constant current convertor. This is the Boucherot effect.

Now at resonance, $\omega^2 LC = 1$, so the load current can be written in various ways as:—

$$I = \frac{V}{j\omega L} = -\frac{jV}{\omega L} = -jV\omega C = -jV\sqrt{\frac{C}{L}}$$

In Fig. 3 the load current is:—

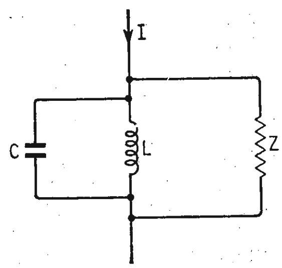
$$I = jV\omega C = \frac{jV}{\omega L} = jV\frac{C}{L}$$

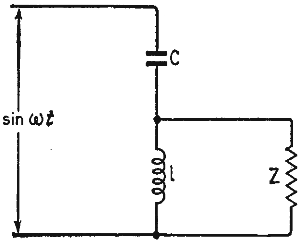
It has been assumed above that the components L and C are pure lossless reactances and that they resonate at the exact frequency of the supply voltage. When making use of the Boucherot effect in practice we must consider what will happen if there are losses in the reactive components or if the frequency varies.

Frequency Variations.—Suppose in the circuit of Fig. 1 that the applied frequency is above resonance. Then the value of the capacitor would have to be reduced to restore resonance. Consider the actual capacitor to be formed from two capacitors in parallel as shown in Fig. 4, where one capacitor C_1 resonates with L at the applied frequency. Then L and C_1 form a Boucherot constant current circuit having C_2 in parallel with Z as its load. The capacitor C_2 therefore diverts some of the constant current $V/j\omega L$ from the load Z.

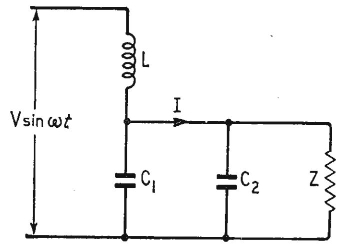
Similarly, a reduction of frequency has the same effect on the circuit of Fig. 1 as the addition of an

Fig. 2. The circuit of Fig. 1 may, by application of Norton's theorem, be considered in this form.





Left: Fig. 3. Inverted form of circuit in Fig. 1.



Right: Fig. 4. At frequencies higher than the resonant frequency of LC, C splits and constant current is shared between C_2 and Z.

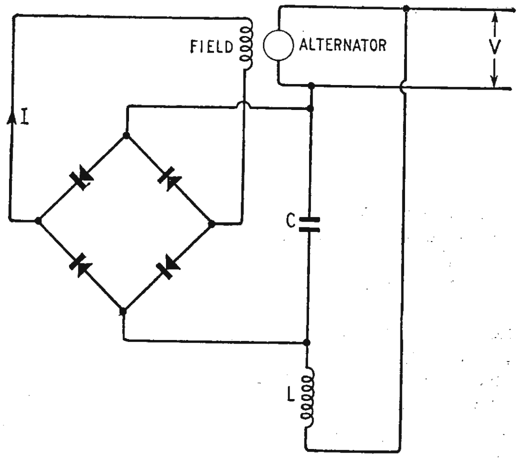


Fig. 5. Boucherot circuit provides constant voltage to field current rectifier at low speeds.

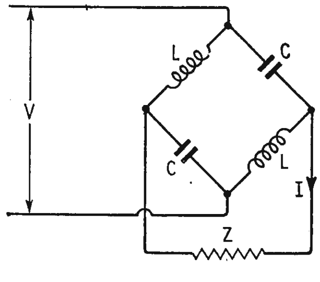
inductor in parallel with the load Z. A similar approach can be used with the circuit of Fig. 3.

If the inductor has appreciable losses, it is most convenient to consider these as caused by a resistor in parallel with the inductor. The effect on the operation of the circuit of Fig. 3 is then the same as would be caused by the use of a perfect Boucherot circuit, but with the loss resistor added across the load impedance Z. This also applies to Fig. 1, but here the short circuit current is also changed by the addition of a loss resistor across the inductance.

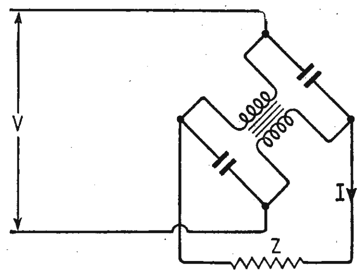
The supply current for the circuit shown in Fig. 3 is: $I_s = V(Z + j\omega L)/\omega^2 L^2$, so the magnitude of Z must not be made too large if excessive supply currents and component voltages are to be avoided. If the load Z is disconnected, we are left with a series resonant circuit across the supply and this is, of course, a short circuit. When the load has a high impedance, the Boucherot circuit produces a high voltage to force the correct load current to flow.

Applications.—One practical application of the Boucherot circuit has made use of this fact. The impedance of a normal semiconductor rectifier is very high when the applied voltage is low. Consequently little load current can flow in the load connected to the rectifier at low applied voltages and the linearity between a.c. input and d.c. output is poor. However, if the rectifier is supplied from a Boucherot circuit it is forced to take a current proportional to the alternating supply voltage regardless of the rectifier impedance or of the voltage value, and this current must flow through the load provided that the reverse resistance of the rectifier is reasonably high. An arrangement of this type is shown in Fig. 5. When equipment must be supplied from an engine-driven generator, the field current required for the generator can be obtained from the output if a d.c. generator is used. If, in order to make it possible to use transformers to obtain a number of different voltages, an alternator is used instead of a d.c. generator, the d.c. field current can be obtained by rectifying the output. However, when starting up the alternator, there is not sufficient output voltage present to overcome the high impedance of the rectifier. A Boucherot circuit used as shown in Fig. 5 can force field current through the rectifier and so cause the alternator output voltage to build up from its normal low residual value.

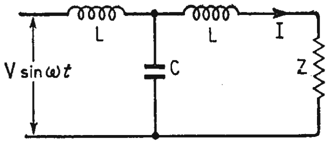
The Boucherot circuit has been widely used for lighting supplies where long chains of lamps are in series[2]. The advantage of such a connection is that the length and size of cable needed to supply the lamps is much less than if they were all in parallel, and voltage drop along the cables is less of a problem. A Boucherot circuit can supply the lamps with constant current regardless of their impedance variations. If a particular lamp fails it can be shorted out by an automatic circuit, and the Boucherot circuit will continue to maintain constant the current in the rest of the lamps. Airfield lighting supplies[3] are often arranged in this way, a bridge form of the Boucherot circuit (named the "monocyclic square" by Steinmetz) normally being used. This is shown in Fig. 6 where it can be seen that each half of the



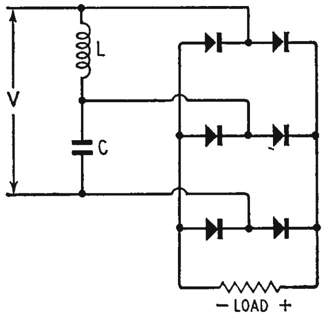
Left: Fig. 6. Steinmetz monocylic square reduces phase shift between supply voltage and current.



Right: Fig. 7. Coupling between coils equalizes voltages across bridge components.



Left: Fig. 8. Constant current to constant voltage converter. The similarity to a low-pass filter is worth noting.



Right: Fig. 9. Single to three-phase converter avoids need for large smoothing circuit.

bridge takes the form of Fig. 1, one half being inverted with respect to the other. The advantage of the bridge connection is that the supply current is more nearly in phase with the supply voltage than is the case for a half circuit, but Boucherot pointed out that the components must be accurately matched if the supply is not to be effectively short circuited. In order to avoid large differences between the voltages across various components, it is necessary to use the arrangement of Fig. 7, in which the transformer is of 1 : 1 ratio, or to detune the circuit slightly.

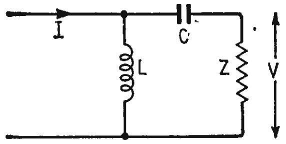
While the bridge circuit is often used, there is another way of making the input current in phase with the input voltage when the load is resistive. This is shown in Fig. 8. Here the load current is still equal to $V/j\omega L$, but the input current is now equal to $V\omega^2C^2Z$.

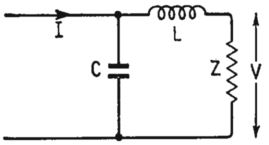
In industrial electronic equipment the large ripple voltage produced by a single phase full wave rectifier is often a nuisance. It is desirable to avoid the use of high value smoothing capacitors in order to save space and to improve reliability. In addition, when varying alternating voltages have to be rectified in servo and regulator applications, the slow response of a smoothing circuit to variations of the amplitude of the alternating voltage input is a severe disadvantage. A method of using the Boucherot circuit to reduce the ripple produced and therefore the amount of smoothing required is shown in Fig. 9. Here a three-phase full-wave rectifier has one input terminal supplied from a Boucherot circuit and the other two terminals connected to the main single phase supply. Provided that the load resistance has a suitable value, a balanced three-phase voltage system is produced at the terminals of the rectifier, and the rectifier only produces a low amplitude sixth harmonic ripple. Little smoothing is required on the d.c. side of the rectifier to remove this high frequency, low amplitude ripple, and it is possible for the d.c. output to respond rapidly to variations of the alternating input voltage amplitude. These characteristics make the circuit a useful one in industrial electronics. It should be noted, however, that if the supply voltage is reduced to a low value the percentage ripple will increase. The reason for this is that as the voltage decreases the impedance of the rectifier increases and in attempting to force current through the increased impedance the Boucherot circuit unbalances the three-phase voltage system and so introduces second harmonic ripple.

Constant Voltage Circuits

The dual of the Boucherot circuit is shown in Fig. 10. This can produce a constant voltage across the load when supplied from a constant current, constant frequency source. To show this, if the load Z in Fig. 10 is open circuited, the output voltage is $Ij\omega L$, where I is the supply current. Now Thevenin's theorem states that this open circuit voltage is effectively applied to the load in series with the internal impedance of the circuit. Since in this case the supply is from a constant current source which therefore has an infinite internal impedance, the impedance looking in at the output terminals is simply formed from L in series with C, the infinite supply impedance being across C. Now if L and C are in series resonance at the supply frequency, the internal impedance seen from the output terminals is zero, and the whole of the open circuit voltage $Ij\omega L$ appears across the load Z regardless of the magnitude or nature of the latter. Similarly, in Fig. 11 the load voltage is $I/j\omega C$ (which is equal to $-Ij\omega L$ since the circuit is resonant) and it is again independent of Z. In both cases the load voltage is phase shifted 90° to the supply current I.

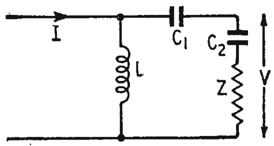
Right: Fig. 10. Dual of Fig. 1 giving constant voltage.





Left: Fig. 11. Inversion of circuit in Fig. 10.

Right: Fig. 12. At frequencies below resonance, C splits, and constant voltage is developed across C_2 and Z in series.



Frequency Variations.—The effect of frequency changes on this constant current to constant voltage convertor can be visualized as follows: suppose, in the circuit of Fig. 10, that the applied frequency is below resonance. Then the capacitor would have to be increased in value to restore resonance. Consider the capacitor to be formed from two capacitors in series as shown in Fig. 12, where one capacitor C1 resonates with L at the applied frequency. Then the voltage across C_2 and Z in series will be maintained constant at $Ij\omega L$ by the inverse Boucherot circuit formed from L and C_1. Similarly an increase of supply frequency has the same effect on the circuit of Fig. 10 as the addition of an inductor in series with the load Z. Similar procedures can be used with the circuit of Fig. 11.

If the inductor in the inverse Boucherot circuit has appreciable losses, they should be considered as being due to a series resistor. The effect on the circuit shown in Fig. 11 is then the same as the addition of the loss resistor in series with the load Z. The same applies to Fig. 10, but here the open circuit voltage is also changed by the addition of a loss resistor in series with the inductor.

The supply voltage for the circuit of Fig. 10 is $I(1 + j\omega CZ)/\omega^2C^2Z$, so Z must not be too small if excessive voltages and currents are to be avoided. The magnitude of the supply voltage is $I\omega L(\omega^2L^2/Z^2 + 1)^{\frac{1}{2}}$. The square root is plotted in Fig. 13 against $Z/\omega L$. It will be seen that if Z is resistive and it is less than ωL the voltage required increases rapidly, and this is one limitation of the circuit.

Applications.—Typically, this scheme can be employed as shown in Fig. 14. Here a pentode is used as the constant current generator, with an inductor in its anode circuit. Effectively across the inductor are a capacitor and the load in series, the power supply being assumed to have a low internal impedance. The capacitor resonates with the inductance at the frequency of the voltage applied to the grid of the pentode. In such a circuit, the load voltage is maintained almost constant as the load impedance is varied, provided that the load impedance is not so low that excessive currents and voltages would be required. In a practical arrangement of this type there is a slight slope in the constant load voltage region due to imperfections of the components and constant current generator used. Even if the load is rather non-linear, the load voltage tends to be maintained constant and sinusoidal. Such arrangements are therefore useful where a constant frequency current generator such as a current transformer, a pentode or a transistor is required to provide a constant sinusoidal voltage across a variable or non-linear load. Intermediate frequency amplifiers employing this method of coupling automatically possess this property, although the fact tends to be overlooked because the main interest is in the selectivity of the circuit rather than in what happens at resonance. Another circuit which can convert a constant current to a constant voltage is shown in Fig. 8.

The similarity of some of the circuits which have been given here, for example Fig. 8, to some of the filters and transmission lines used in radio and communications circuits is worth noting. When such filters are operated at or near resonance, it is useful to remember that they can exhibit the Boucherot effect in addition to their textbook properties.

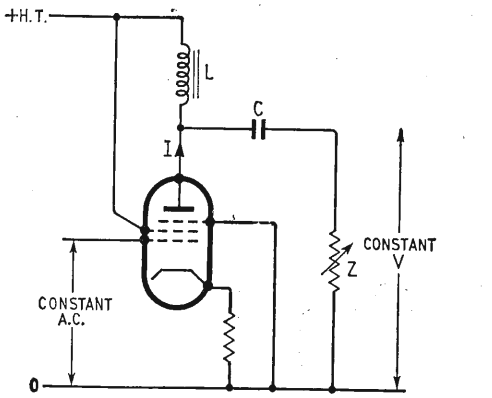


Fig. 14. One application of constant voltage circuit. V remains constant in spite of variations in Z.

An important point in these days of wide application of negative impedances is that a negative impedance in series with a positive impedance of equal but opposite magnitude will exhibit the Boucherot effect[4]. If the two impedances are resistive, the Boucherot effect will be obtained at all frequencies, and it is no longer a resonant effect.

One reason why the Boucherot effect tends to appear more in power applications practice than in electronics at present is that in a power system the internal impedance of the alternating supply is very low, so that it can be considered as a constant voltage source. In electronics this is rarely the case. However, constant current sources such as pentodes and transistors are common enough, and so are non-linear loads. For these reasons, the properties of the dual Boucherot circuits of Fig. 10 and Fig. 11 could be exploited more widely than they are in electronics for fixed frequency applications. In addition, with the appearance of negative resistance elements such as tunnel diodes, there are attractive possibilities of combining positive and negative resistances to obtain a frequency independent inverse Boucherot effect.

A phenomenon such as the Boucherot effect which has found many applications and potentially has many more should certainly not be ignored.

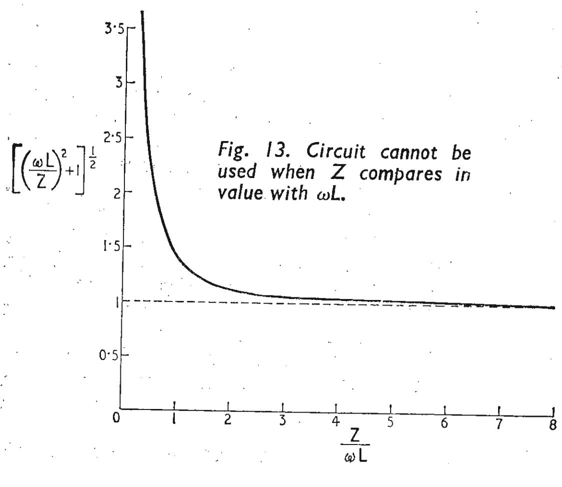


Fig. 13. Circuit cannot be used when Z compares in value with ωL.

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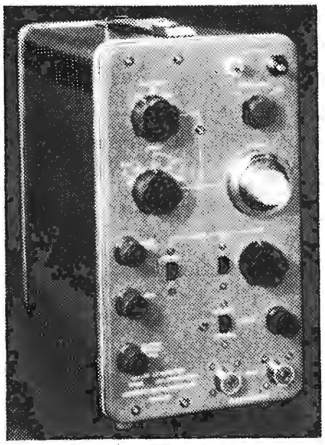
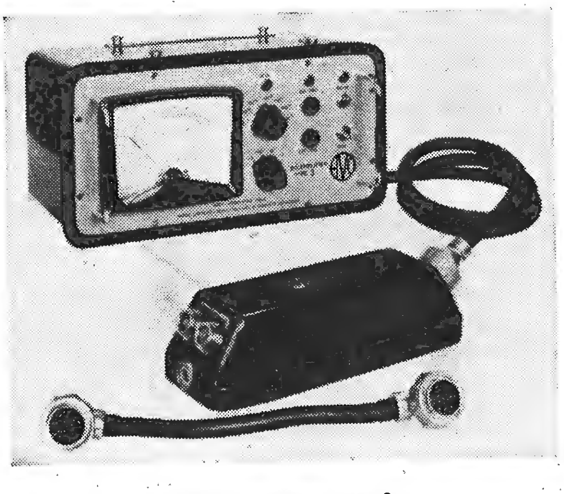


Marconi Instruments delay generator TF1415.

indicating the position of the start of the delayed sweep. Full details are obtainable from Marconi Instruments Ltd., Longacres, St. Albans.

D.C. Amplifier

THE mains-operated d.c. amplifier, Type 3, made by Avo Ltd., 92-96 Vauxhall Bridge Road, London, S.W.1, is suitable for ionization chamber measurements and



Avo d.c. amplifier, type 3.

other applications involving the measurement of currents in the range 3×10^{-13} to 10^{-6}A. The equipment comprises a transistor main amplifier and indicating instrument and an electrometer valve head amplifier which incorporates range-selecting feedback resistors remotely controlled by an electrically operated rotary switch from the main unit. In addition to the direct-current-measuring head there is a current-integrating head, incorporating feedback capacitors from 10 pF to 10μF in six steps and a specialized head with two terminals across which any derived feedback impedance can be placed.

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Portable Pen Recorder

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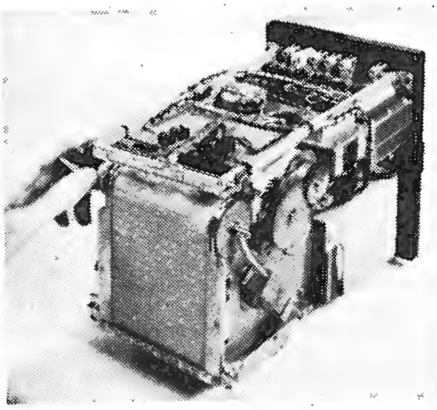


Mervyn portable pen recorder.

of waxed paper which requires no ink. Maximum sensitivity is 5mV for full-scale deflection, and a drift corrected amplifier is employed. Further information may be obtained from Mervyn Instruments, St. John's, Woking, Surrey.

Solder-cooled Transistors

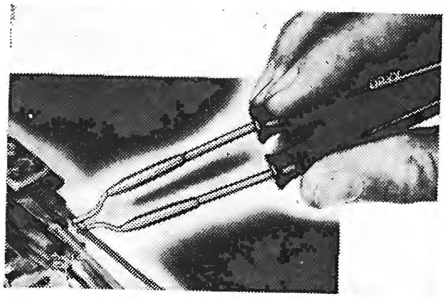
THE idea of solder cooling anything will not receive a sympathetic audience from those of us who have recently burned our fingers; but Enthoven Solders have turned the malleability of solder—and its heat conductivity—to advantage with thin preformed solder-alloy washers for aiding the heat transfer between transistors and their heat sinks. The soft washer of solder is "squeezed" between the transistor and its mounting plate and takes up irregularities in both transistor surface and plate, assuring good thermal contact. Where electrical insulation, by a mica washer, for instance, is required, the preform helps to eliminate fracturing of the mica. Typical temperature measurements quoted are: direct mounting; transistor 33°C, sink 29°C: silicone grease; transistor 33°C, sink 31.5°C: Enthoven preform; transistor 30°C, sink 30°C: Mica washer; transistor 33°C, sink 29°C: Enthoven preform and mica; transistor, 31.5°C, sink 30.75°C.

Enthoven Solders Ltd., Dominion Buildings, South Place, London, E.C.2.

Soldering Tweezers

THE soldering of leads to valve holders, switch and relay contacts can be difficult in closely-packed assemblies because the space available often precludes application of the iron to both sides of the contact should it be found hard to "wet" with solder. A new double iron marketed by W. Greenwood Electronic, under the name "Oryx," has been found useful in these situations: consisting of two 7-W irons with permanent nickel bits, the assembly is made up in tweezer form so that an iron may be applied to both sides of the work simultaneously. The tweezer iron, too, is useful when working on small printed-wiring boards as it enables component wires to be heated from both sides simultaneously, easing withdrawal with the minimum of disturbance to surrounding components.

In practice the iron was found easy to handle and an



"Oryx" soldering tweezers.

improvement over a single iron of similar rating; but, as with any unfamiliar tool, a little practice is necessary to achieve best results and avoid fatigue.

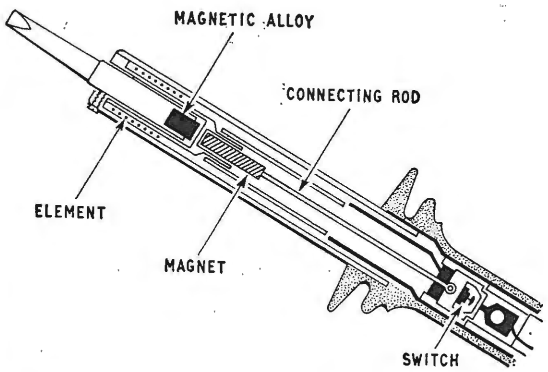
An adaptation of the iron is available as a wire-stripper, where the bits are replaced by twin heated cutters.

W. Greenwood Electronic Ltd., 677, Finchley Road, London, N.W.2.

Temperature-controlled Soldering Irons

THE Curie point, at which a magnetic material loses its magnetic properties, is the nub of the thermostat in the Weller range of heat-controlled soldering irons. The "Magnastat" control uses a cylinder of special nickel-iron alloy with a chosen Curie point: this cylinder is mounted in the end of the bit remote from the soldering tip. In the body of the iron is a permanent magnet attached to a microswitch which controls the power to the element. When the iron is cold the magnet is attracted to the alloy in the bit, so switching on the power: as soon as the iron reaches working temperature the alloy loses its properties allowing the spring in the switch assembly to open the contacts.

The Magnastat principle is applied to irons from 40W (24V only) to 160W (115 to 240V): included in the range is a 70W type with an insulated bit proof-tested at 3.5kV for working on live equipment. The 55W iron weighs only 3oz and a range of replaceable bits is available for all types, giving a choice of tip temperatures between 410 and 750°F.



"Magnastat" control system of Weller soldering iron.

British agents for Weller are Elstone Electronics Limited, Hereford House, North Court, Vicar Lane, Leeds 2.

Plugs, Sockets and Transistor Holders

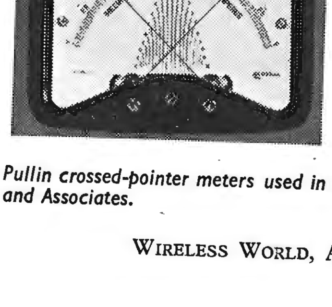
MADE with four to ten ways, the Electromethods series M, BMA and BHM miniature plugs and sockets are moulded in Melamine and employ gold-plated contacting surfaces. Requiring a mounting hole only 0.5in in diameter, ratings are:—type M 350V, 7.5A; type BMA 250V, 10A; type BHM (hermetically-sealed plugs) 350V, 5A. Type BHM is soldered into a bulkhead or metal container. Accessories available include aluminium hoods for the free members (as an alternative to Melamine), locking clips, cable clamps and protective pin shrouds.

The M9T transistor holder is based on these plugs and sockets and would appear to be most useful for circuit-development work. The plug is provided with four contacts: the transistor is mounted in a well in the plug body and has the cover screwed over it. The transistor is thus protected against mechanical damage, breakage of leads and deterioration caused by frequent resoldering in experimental use.

Electro Methods Ltd., Electrical Connector Division, Hitchin Street, Biggleswade, Beds.

Crossed-pointer Meters

THE use of crossed pointers in an indicating meter is well known in aircraft instruments, and their advantages of convenience and space-saving are now extended to general-purpose instruments by Measuring Instruments (Pullin) Ltd., Electrin Works, Winchester Street, Acton, London, W.3. The movements can be a.c. or d.c., moving-iron or moving coil, and incorporate a centre-scale, on which the pointer intersections indicate ratios or products of the two readings.



One of the Pullin crossed-pointer meters used in a computer by A.M. Gear and Associates.

The Darlington Connection

COMPOUND CIRCUIT WITH α VERY NEARLY UNITY

By O. GREITER

A THERMIONIC valve's current gain is nearly infinite in the common-cathode mode of operation. Few readers will remember just how many microamps of grid current are taken to control 100mA of anode current in a 6L6 or an EL34, although there is a difference and it can affect the performance of a driver stage. An ideal transistor, too, would have infinite current gain in the common-emitter mode. Holes injected at the emitter should all diffuse through to the collector junction and there should be no current in the base lead at all. Practical transistors are by no means ideal and the main task of the circuit designer is to approximate, with imperfect devices, the results he could easily achieve if the devices were ideal.

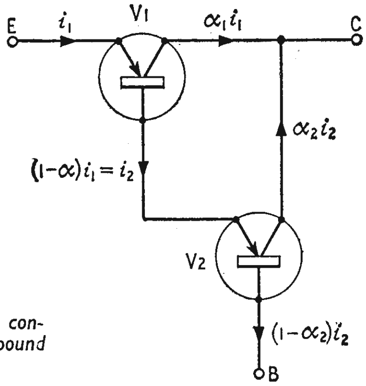


Fig. 1. Darlington's connection forms compound transistor.

A very important technique which is becoming even more widely used is the form known variously as the Darlington connection, the compound connection or the composite transistor. The basic form of the circuit is shown in Fig. 1 and the standard analysis begins with the assumption that a current i_1 is fed into the emitter of V1. As a result a current $\alpha_1 i_1$ flows out at the collector and thus a current $(1-\alpha_1)i_1$ flows out at the base. This current we can call i_2 and it flows into the emitter of V2.

The total current flowing out at C will be $(\alpha_1 i_1 + \alpha_2 i_2)$ and the current flowing out at B will be $(1-\alpha_2)i_2$. If we call these currents I_C and I_B we have then,

$$I_C = (\alpha_1 i_1 + \alpha_2 i_2) = \alpha_1 i_1 + \alpha_2(1-\alpha_1)i_1 = (\alpha_1 + \alpha_2 - \alpha_1\alpha_2)i_1$$

$$I_B = (1-\alpha_2)i_2 = (1-\alpha_1)(1-\alpha_2)i_1$$

This is the behaviour which would be obtained from a single transistor having a current gain of

$$\alpha = \alpha_1 + \alpha_2 - \alpha_1\alpha_2 = 1-(1-\alpha_1)(1-\alpha_2)$$

When we take $\alpha_1 = \alpha_2 = 0.98$ we get

$$\alpha = 1-(0.02)^2 = 0.999{,}6$$

so that $\alpha' = 1/(1-\alpha) = 2{,}500$

and if $\alpha_1 = 0.9$, $\alpha_2 = 0.98$ we have

$$\alpha = 1-(0.1)(0.02) = 0.998$$

It will be seen that α remains very near unity even when the current gain of one transistor has fallen substantially. The reason is that if V1 has a low gain and the base current is substantial this current is not lost but is brought round to terminal C through V2, provided that V2 has high gain. This is a point of considerable importance.

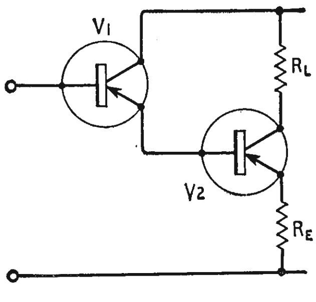
The compound pair is usually used in the common emitter mode. The typical form is shown in Fig. 2 and the current gain of the composite transistor is

$$\begin{aligned}\alpha_c' &= (\alpha_1 + \alpha_2 - \alpha_1\alpha_2)/(1-\alpha_1)(1-\alpha_2)\\ &= [1-(1-\alpha_1)(1-\alpha_2)]/(1-\alpha_1)(1-\alpha_2)\\ &= [1/(1-\alpha_1)(1-\alpha_2)] - 1\end{aligned}$$

Clearly this is a very large number, perhaps 500 or even 5,000. It will be sensitive to changes in α_1 and α_2 so that linearity will depend on our making full use of the high linearity of the emitter/collector current relationship by the feedback resistor R_E. This brings out an important difference between this circuit and the common-emitter common-collector pair shown in Fig. 3. For this circuit the current gain is $\alpha_1/(1-\alpha_1)(1-\alpha_2)$, which is also a very large number and it might seem that there is gain enough to make the rather small difference unimportant. With the circuit of Fig. 3, however, any current lost at the base of V1 is not recovered in the load resistor R_L and, so far as any linearizing effect against changes in α_1 is concerned, we have only the gain in the transistor V1.

Against this we must set an important advantage in the use of the Fig. 3 circuit. With the compound connection the voltage applied to the collector of the driver transistor is equal to the voltage applied to the collector of the main transistor and the emitter current of the driver transistor must be the base current of the main transistor. Let us take the OC26 as the main transistor and let us demand a peak current of 3A. For a minimum current gain we have $\bar{\alpha}' = 15$, so that we need a base current of 100mA peak. The driver transistor for a Class-A stage will then need to operate at about 50mA. With a 12-V supply we can hardly expect to have less than some 10V across the transistor and the driver will be dissipating 500mW and

Fig. 2. Compound common-emitter pair with collectors joined.



the OC26 will be slightly over its limit under these conditions.

It is most unlikely that we shall need more than about a 2-V excursion at the base of the main transistor and, if the circuit of Fig. 3 is used, we might limit the collector-to-emitter voltage of the driver transistor to some 4V with perfect safety. The dissipation then drops to some 200mW. As it happens this, as a region of power dissipation, is a rather awkward one, for a dissipation of 500mW is too much for the small transistor and is very low down on the characteristics of the power transistor. As the range of devices is increasing (and prices falling), the compound connection appears to be gaining in popularity. The basic circuit shown in Fig. 3 remains important, however, especially as the addition of resistance in the collector lead of the driver transistor can be used to make this transistor inherently safe against thermal runaway.

The npn-pnp combination shown in Fig. 4 does not appear to be described in the ordinary text-

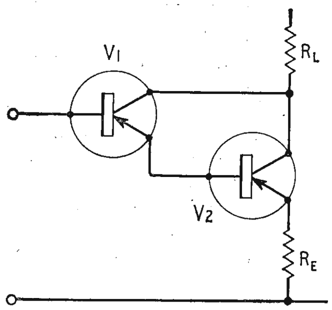


Fig. 3. Here V1 has its collector connected to supply and acts as plain emitter follower driving V2.

books. If a current i_1 flows into the emitter of V1 we get a current $\alpha_1 i_1$ at the collector and a current $(1 - \alpha_1)\, i_1$ at the base. In consequence we have a current at B of $(1 - \alpha_1)\,(1 - \alpha_2)\, i_1$ and a current flowing back towards E of $\alpha_2(1 - \alpha_1)i_1$. The net current at E is therefore $[1 - \alpha_2\,(1 - \alpha_1)]i_1$ and the current gain from E to C is $\alpha_1/(1 - \alpha_2 + \alpha_1\alpha_2)$. This expression can be put equal to a current gain of α for the complete unit and then we can write:

$$1/\alpha = (1/\alpha_1) + \alpha_2 - (\alpha_2/\alpha_1)$$

Let us take, as before, $\alpha_1 = \alpha_2 = 0.98$

$$1/\alpha = 1.020{,}4 + 0.98 - 1 = 1.000{,}4$$

so that $\alpha = 0.999{,}6$. It is seen that this configuration also gives the very high composite current gain. Here, however, the voltage across the driver transistor (V2) is low and would be some 200 to 600mV for the OC26. The lower voltage is that at which V1 is passing no base current and it would appear that the OC139, which bottoms at $v_{ce} = 175$mV with 7.5mA collector current will just about fit the bill. It will then operate normally around 0.5V, 50mA, a dissipation of only 25mW, which allows the transistor to be used up to an ambient temperature of 60°C.

This discussion of the compound connection leaves us with a rather limited knowledge of the overall behaviour of the composite transistor. It is possible to generate a set of performance curves by combining the characteristics of the two transistors point by point. If this is sufficient it is probably not too tedious an operation because we need take only three points, the centre of the range, near cut off and at full drive in the linear region just below bottoming. Indeed, when the device is to be used (as it frequently is) in a Class-B stage, we may content ourselves with two points only. Observe, however, that we are on the horns of a dilemma. If this information is sufficient for our work with a composite transistor it is sufficient when we have a simple transistor: if we need a full set of parameters with a simple transistor we need them with a composite transistor. We know that in practice we need the full set of parameters only occasionally because we have learned that with particular operating conditions we can neglect some factors. An example of this is the way in which we usually neglect the anode impedance of a pentode valve and rarely know what value of amplification factor it has. The moment of truth in this connection comes when we come to operate a so-called starved amplifier and find that here these two factors must be kept separate and not combined in a single mutual-conductance term.

The derivation of the full set of transistor parameters is inevitably a tedious operation which is most easily carried out with numbers. Here, of course, we must work with symbols but the reader is recommended to follow the symbolic procedure with numerical values rather than to insert numbers at the end.

For purposes of analysis we can re-draw the circuit of Fig. 1 in the form shown in Fig. 5, re-numbering the transistors. For V1 we can write the standard hybrid h equations:

$$v_1 = h_{11}'I_1 + h_{12}'v_2$$
$$I_2 = h_{21}'I_1 + h_{22}'v_2$$

The h parameters used here are those for the common-collector mode of operation and are related to the common-emitter parameters normally available by the equations

$$h_{11,c} = h_{11,e}$$
$$h_{12,c} = 1 - h_{12,e}$$
$$h_{21,c} = -(1 + h_{21,e})$$
$$h_{22,c} = h_{22,e}$$

The prime (′) has been dropped because it serves merely to indicate V1, and we shall use h'' for V2. We shall also need the determinant

$$h_{11,c}\, h_{22,c} - h_{12,c}\, h_{21,c} = \Delta h_c = 1 + h_{21,e} + \Delta h_e - h_{22,e}$$

In terms of the T-network parameters an approximation is

$$h_{11,c} = r_b + r_e/(1-\alpha)$$
$$h_{12,c} = 1$$
$$h_{21,c} = -1/(1-\alpha)$$
$$h_{22,c} = 1/r_c(1-\alpha)$$

The reader who is concerned with greater precision will find fuller expressions given in Shea's *Transistor Circuit Engineering* (John Wiley, 1957).

In order to analyse the circuit of Fig. 5 it is convenient to rearrange the basic equations above in the form:—

$$v_1 = -(\Delta h'/h_{21}')\, v_2 + (h_{11}'/h_{21}')I_2$$
$$I_1 = -(h_{22}'/h_{21}')\, v_2 + (1/h_{21}')\, I_2$$

and then we can write, for the second transistor,

$$v_2 = -(\Delta h''/h_{21}'')\, v_3 + (h_{11}''/h_{21}'')\, I_3$$
$$-I_2 = -(h_{22}''/h_{21}'')\, v_3 + (1/h_{21}'')\, I_3$$

It now becomes merely a matter of substituting the expressions for v_2 and I_2 from the second set of equations into the first set and re-arranging to get:—

$$v_1 = [(\Delta h'\,\Delta h''/h_{21}'h_{21}'') + h_{11}'\,h_{22}''/h_{21}\,h_{21}'']v_3 - [(\Delta h'\,h_{11}'' + h_{11}')/h_{21}'\,h_{21}'']I_3$$

$$I_1 = [(\Delta h''\,h_{22}' + h_{22}'')/h_{21}'\,h_{21}'']\,v_3 - [(h_{22}''\,h_{11}' + 1)/h_{21}'\,h_{21}'']\,I_3$$

At this point it is essential to introduce some new abbreviations and we can write these equations as:—

$$v_1 = A_{11}v_3 - A_{13}I_3$$
$$I_1 = A_{31}v_3 - A_{33}I_3$$

from which we can obtain, by rearrangement:—

$$v_1 = (A_{13}/A_{33})\,I_1 + (\Delta A/A_{33})v_3$$
$$I_3 = -(1/A_{33})I_1 + (A_{31}/A_{33})\,v_3$$

These four coefficients are the common-collector hybrid-h parameters of the composite transistor and they give us, as the approximate common-emitter hybrid-h parameters

$$h_{11,de} = A_{13}/A_{33}$$
$$h_{12,de} = 1 - \Delta A/A_{33}$$
$$h_{21,de} = 1/A_{33}$$
$$h_{22,de} = A_{31}/A_{33}$$

Let us now plunge into arithmetic assuming two similar transistors having $h_{11,e}' = 1\mathrm{k}\Omega$, $h_{12,e}' = 10^{-3}$, $h_{21,e}'=49$ and $h_{22,e}'=10^{-4}$. (This is a well-rounded OC71.)

We first produce the approximate common-collector parameters:—

$h_{11,c}=1{,}000$, $h_{12,c}=0.999$, $h_{21,c}=-50$, $h_{22,c}=10^{-4}$ and $\Delta h_c = 0.1 + 50 \times 0.999 \approx 50$.

From these we can write down

$$A_{11} = (1/2{,}500)\,(50.50 + 1{,}000.10^{-4}) \approx 1$$
$$A_{13} = (1/2{,}500)\,(50.1{,}000 + 1{,}000) \approx 20$$
$$A_{31} = (1/2{,}500)\,(50.10^{-4} + 10^{-4}) \approx 1/500{,}000$$
$$A_{33} = (1/2{,}500)\,(1.1) \approx 1/2{,}500$$

so that $\Delta A = (1/2{,}500) - (20/500{,}000) \approx 1/2{,}500$

and thus $h_{11,e\,(\mathrm{comp})} = 20.2{,}500 = 50{,}000$

$$h_{12,e\,(\mathrm{comp})} = 1-1$$
$$h_{21,e\,(\mathrm{comp})} = 2{,}500$$
$$h_{22,e\,(\mathrm{comp})} = 1/200$$

First of all we must look at $h_{12,e\,(\mathrm{comp})}$. We have here the small difference between two quantities, each of which is nearly unity and in making our approximation we have "thrown the baby out with the bath water." We can look at matters in a slightly

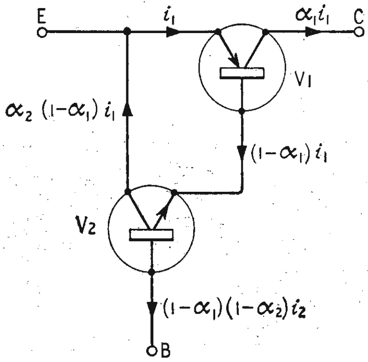


Fig. 4. npn-pnp compound pair of very high current gain.

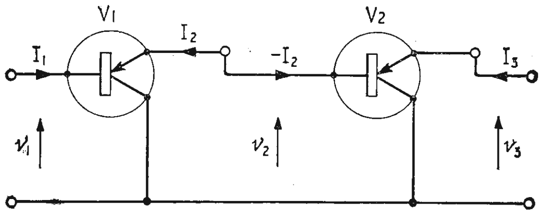


Fig. 5. Darlington's connection redrawn for purposes of analysis.

different way by going back and working the correct value out in detail. The term we want is $h_{12,e}$ and we have $h_{12,e} = 1-h_{12,c}$. Thus the small difference must appear in ΔA and, in full:—

$$\Delta A = [(h'\,\Delta h''+h_{11}'h_{22}'')\,(1+h_{11}'\,h_{22}'') - (\Delta h''\,h_{22}' + h_{22}'')\,(\Delta h'\,h_{11}''+h_{11}')]/(h_{21}'h_{21}'')^2$$
$$\approx [(2{,}500)-(50.10^{-4}.50.10^{3})]/(2{,}500)^2$$
$$\approx 2{,}250/(2{,}500)^2$$

This gives us $\Delta A/A_{33} = 2{,}250/2{,}500$ and thus $h_{12,e\,(\mathrm{comp})} = 0.9$. The result is still an approximation but it serves to illustrate how the feedback term has now become substantial. It will be clear that what is happening is that the feedback in one transistor is being multiplied by the gain in two and is a much more serious matter now. The term $h_{11,e\,(\mathrm{comp})}$ has been raised by a similar mechanism, for the resistance of the base-to-emitter diode of the second transistor appears as emitter-feedback resistance for the first. The high internal feedback will be seen to have reduced the output impedance. This analysis brings out a very important point; because there is so much feedback through the transistors the input impedance is much more easily affected by the load impedance. The high input impedances which the simple theory predicts are not normally obtained and this is disconcerting unless the engineer is aware that he is working with approximations which are not valid.

Analysis of the circuit of Fig. 4 requires only that we should use the common-emitter parameters for V2 in place of the common-collector parameter. If we do this we find that, as we might expect, the feedback term remains very nearly that of a single transistor while $h_{11,de}$ is very high. The complementary composite structure is thus apparently much superior to the conventional form.

The results obtained here owe a good deal to the analysis by Ghandhi in *I.R.E. Transactions on Circuit Theory* (Sept. 1957) in which matrix algebra is used openly to arrive at some rather more elegant forms. In terms of the common-collector parameters we find there, for the common-emitter composite:—

$$H_{11} = (h_{11}' + \Delta h'\,h_{11}'')/(1 + h_{11}''\,h_{22}')$$
$$H_{12} = 1 - h_{12}'\,h_{12}''/(1 + h_{11}''\,h_{22}')$$
$$H_{21} = [h_{21}'\,h_{21}''/(1 + h_{11}''\,h_{22}')] - 1$$
$$H_{22} = (h_{22}'' + \Delta h''\,h_{22}')/(1 + h_{11}'\,h_{22}')$$

and for the common-base composite:—

$$H_{11} = (h_{11}' + \Delta h'\,h_{11}'')/(\Delta h'\,\Delta h'' + h_{11}'\,h_{22}'')$$
$$H_{12} = 1 - h_{21}'\,h_{21}''/(\Delta h'\,\Delta h'' + h_{11}'\,h_{22}'')$$
$$H_{21} = [h_{12}'\,h_{12}''/(\Delta h'\,\Delta h'' + h_{11}'\,h_{22}'')] - 1$$
$$H_{22} = (h_{22}'' + \Delta h''\,h_{22}')/(\Delta h'\,\Delta h'' + h_{11}'\,h_{22}'').$$

UNBIASED

By "FREE GRID"

Why "Free Grid"?

THE editor tells me that on more than one occasion he has been asked to explain why I write under the pen name of "Free Grid," and why my writings bear the title of "Unbiased." In these transistorized times, the thermionic valve has taken rather a back seat, but all the same, I am sure no reader of *Wireless World* needs telling that, when the control grid of a valve is "free," it is in an unbiased state. In fact, the grid is not only in an unbiased state but is also in a completely uncontrolled condition, and that applies also to the valve to which the grid belongs.

Thirty-two years ago when the then editor, who is now the Managing Editor, invited me to write for *Wireless World*, the home-construction phase of wireless development was at its height, and details of many sets by various designers appeared in the pages of *W.W.* All those designs were basically sound but some were more difficult to build and get going than others; in fact with certain sets, a great deal of skill was needed to get the same first-class results as their designers did.

It was felt that readers who constructed these sets from the published designs needed a champion who was himself a home constructor and who could, therefore, see things from their point of view, and was willing and able to take up the cudgels in their defence against designers, manufacturers of components and even against the editor himself when necessary. In fact, what was wanted was a sort of public relations officer in reverse who would be the spokesman of the aggrieved customers rather than one who held a brief for the company which paid him, as with the ordinary public relations officer.

Owing to the fact that I had a certain amount of Irish blood in me, although not actually an Irishman, I have always been "agin the government," and indeed still am. I always was, still am, and always shall be a person holding strong opinions about everything, and it was felt by the editor that I might make a good "devil's advocate," and be always on the side of the underdog—or in other words my fellow readers—against those set in authority.

The editor agreed that I should be free to express what opinions I liked within the limits set by the laws of libel, and so naturally I wanted my column to be called "uncontrolled" and my pen name to be "Free Grid." Unfortunately, the editor decided that "uncontrolled" was not exactly a technical term like "Free Grid," and that therefore, the title must be "Unbiased."

When I suggested that we could make my suggested title of "uncontrolled" *sound* technical by hellenicizing it into "anarchy," he came down very heavily against my idea and decreed that while I could use my suggested pen name of "Free Grid," he must insist on the title being "Unbiased."

I had another go at him, this time on technical grounds, by pointing out that, although a free grid is undoubtedly an unbiased one, an unbiased grid is not necessarily free of control, for when connected to the negative side of the filament to secure freedom from bias it is very much under control. Also I pointed out to him, that when considered in its non-technical sense the title of "Unbiased" would be apt to make readers think that I was never to express any strong opinions about anything which, to anybody who reads my columns, has proved manifestly untrue. However, all editors have much in common with Draco, and so I might as well have saved my breath.

Digitized Jurisprudence

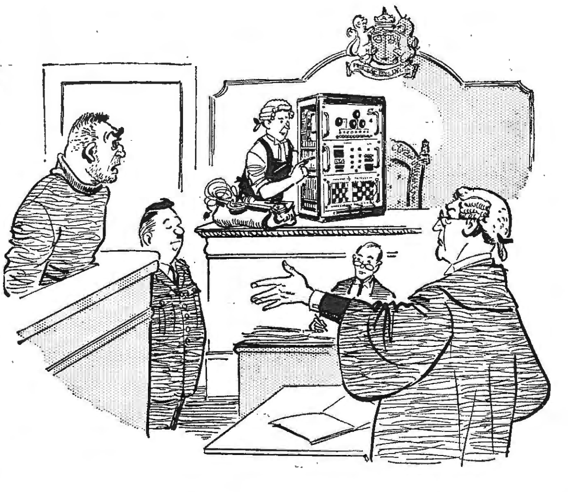
ONE of the most interesting stands at the I.E.A. exhibition held in June was, in my opinion, that of the Ministry of Aviation, largely because it exhibited the Automatic Picture Digitiser. This device will benefit various projects, including—to quote the exact words from one of the Ministry's technical information leaflets—"use of computers for logical processes, e.g., jurisprudence."

Now this obviously means that at some future date an electronic computer will be able to take the place of a jury, an assumption which one of the experts on the Ministry's stand agreed was perfectly correct. A jury is, after all, merely a computer in human form, having the task of processing and assessing information fed to it, and then giving a simple Yes or No answer about the prisoner's guilt. It is, however, a very inefficient and unreliable sort of computer because it is liable to be so swayed by human emotion that it misprocesses the information fed to it.

An electronic computer, however, is devoid of human failings, and, of course, if the defendant's legal advisers thought that it had delivered a wrong verdict, due, for instance to a technical fault in its innards, an immediate appeal would be made to another computer.

Naturally, the place of the judge in courts of the future will be taken by the maintenance man, and if the machine in his care is indeed found to have a fault, he himself will be arraigned before another jurisprudential computer on a charge of neglect of duty, in much the same way as a present-day judge is sometimes virtually on trial before an Appeal Court for alleged faults in his summing up.

To my mind the Ministry of Aviation's experts are not sufficiently far-seeing in this matter of the future application of computers. We all know that, according to modern psychiatric opinion, committing a crime indicates that the perpetrator needs medical attention. Therefore, as I visualize it, after a jurisprudential computer had found a man guilty



"In courts of the future."

he would not be sentenced to time-wasting and quite useless imprisonment. He would be passed at once to a special computer which would process all the digitised information fed to it concerning his mental and bodily health, and thereby arrive at a correct diagnosis, and judicate the necessary therapeutic treatment.

At first, I suppose, an ordinary flesh-and-blood doctor would carry out the treatment indicated by the computer, but in the fullness of time, a special Hippocratic machine will be developed for this purpose. Eventually one can even visualize one of these machines which had failed in its duties, being arraigned before a panel of its fellow machines, charged with "infamous conduct in a professional respect." But that is, I think, looking rather too far ahead.

"All Mod. Con."

I HAVE been greatly interested in some details which the editor has sent to me about the closed-circuit TV and sound intercommunication system which is to be installed in a projected new London hotel. I think, however, the designers of the system are in danger of missing the bus on one or two points.

All rooms are to be equipped for the reception of stereo-sound programmes to be distributed to them from tapes or discs from some central point in the hotel. This apparatus will obviously be available for the distribution of stereo broadcasting when a regular service starts.

I see that programmes are to be supplied to bathrooms but in mono only and not stereo. I think the designers are missing a trick here as it is tape recorders which are needed in bathrooms. Most men—but strangly enoughly, not women—indulge in cacophonic singing when in the bath. If they could only hear themselves, I think they would be cured of this anti-social habit.

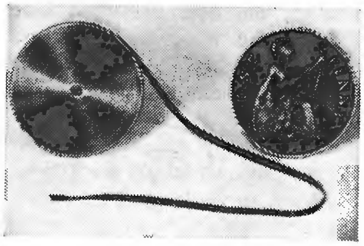
All that is needed is the installation of one of those endless-band tape recorders to be found nowadays on seaside piers and suchlike places. The insertion of sixpence enables you to record your voice and then listen to it before the endless moving band passes on to the wipe-out head when your efforts are mercifully obliterated. The bathroom installation could easily be arranged so that the initial bellow of the bathroom bawler trigged off an acoustically operated switch which set one of these machines in operation.

Another feature of this intercommunication system is that closed-circuit TV is to be installed so that people in their rooms will be able to see who is enquiring for them at the reception desk before they ask for them to be shown up. An excellent idea for those of us who have a bad memory for names but a good one for faces.

TECHNICAL NOTEBOOK

Counting tubes normally indicate the count reached by the position of the glow in relation to a numbered surround on the front panel. If an in-line read-out by Digitron is required ancillary equipment—valves, transistors or discharge tubes—are needed between the Dekatron and the Digitron indicator tube. A modified form of Dekatron, the GCA10G, has been introduced by Ericsson Telephones, of 22, Lincolns Inn Fields, London, W.C.2, which contains ten auxiliary anodes between central anode and the outer ring of cathodes. The additional anodes provide current for the direct operation of a Digitron, and can be used to provide the pre-bias voltage required by the Digitron.

Microtape developed by the American firm Ferrodynamics (English distributors; Electro Techno Dynamics) for ultra-miniature tape recorders is only 0.075 in wide and 0.012 mm thick: the 1¼-in reel shown in the photograph containing 180 feet. The tape has equal thicknesses of polyester base and magnetic coating and it is claimed that a signal-to-noise ratio (for 8.5% total harmonic distortion) of 63dB can be achieved.



The recommended tape tension is 50 gm. Microtape is initially being manufactured in small quantities for use in the development of miniature tape recorders.

New T-R cell incorporating a p-n diode introduced by the Electronic Apparatus Division of Associated Electrical Industries eliminates the need for a "keep-alive" discharge. In operation the diode is pre-pulsed so that it is biased to produce a short circuit across the cell just prior to and during the magnetron transmitter pulse. This attenuates by some 30 dB the power which reaches the receiver and thus protects the receiver input convertor crystal. Most of the magnetron power is thus reflected back towards the input window where it produces a stepped up r.f. voltage. This strikes a discharge which then carries the main power flow from the magnetron to the aerial. In a conventional T-R cell this discharge is struck directly from the transmitted pulse and is also used to protect the receiver (by reflecting power away from it). This means that further protection must be provided against the energy (called spike leakage) which would otherwise reach the receiver in the time interval between the beginning of the transmitter pulse and the formation of the discharge. In a conventional T-R cell such protection is provided by a second or "keep alive" discharge: in the new A.E.I. cell the keep-alive discharge is replaced by the p-n diode, which requires much less power to operate. Faster recovery times than with conventional cells are attainable because additional discharge quenching agent can be used in the cell. This additional quenching agent, and the absence of a keep-alive discharge, also considerably increase the cell life.

Phantom amplifier* provides two-channel working with a single amplifier. Used for stereo, the amplifier consists of a push-pull circuit without phase-splitter: the left and right stereo signals are fed in, one phase-reversed, and they combine to produce a push-pull sum signal, which is extracted in the usual way (the output transformer has a centre-tapped secondary) and a "single-ended" difference signal to which the amplifier behaves as if it were paralleled stages: this signal passes through a transformer connected between h.t. and the centre tap of the push-pull transformer's primary. The secondary of the difference transformer is connected to the centre tap of the sum transformer's secondary so that loads connected between the free end of the difference winding and the sum transformer are energized by left (S-D) and right (S+D) signals, rematrixed from the sum and difference signals.

N. Crowhurst, in *I.R.E. Transactions on Audio* (Vol. AU-9 No. 3) describes a new single-output transformer for this type of amplifier. The transformer can act as loudspeaker filter and, if needed, part of a tone-control network with slopes steeper than those obtainable from a separate single-stage CR or LR network. These facilities employ the leakage inductance of the windings which, because leakage inductance can be described as lack of coupling between windings by the core, behaves as a distortion-free, air-cored inductance.

* See, for example, *Wireless World*, p. 80, Vol. 65, February 1959.

The output transformer employs a single-loop core: on opposite limbs of the core are wound primary and secondary pairs that in themselves are tightly coupled; but which have between them a chosen degree of leakage inductance. The primary coils are operated in push-pull and would provide, without further modification, push-pull outputs at the secondaries at frequencies below the point where the leakage inductance comes into play: also the d.c. core-magnetizing component is cancelled. At high frequencies however, they act as single-ended transformers giving left and right outputs. A third low-impedance centre-tapped winding on a limb at right angles to the other coils also has controlled leakage to the other windings and gives, with the aid of a capacitor forming a low-pass filter with the leakage inductance, a bass-only output. This may be used to energize a separate common-bass speaker, and cancel the partial bass from the "left" and "right secondaries for the use of stereo-only treble speakers or, by inversion of the connections, an output of rising impedance at low frequencies so that greater power may be delivered to two full-range loudspeakers. Leakage inductance between the halves of the separate winding is to be avoided and, to this end, a bifilar style of winding is used.

Tone control may be achieved by variation of the proportions of high and low frequencies used for feedback: the separating action of the transformer is employed in place of an extra network.

Car "road tests" can be carried out in the laboratory by means of a magnetic recording technique demonstrated recently by Fairey. A vehicle is first driven over typical road test surfaces and the vibrations produced at its four wheels measured by suitable transducers and recorded on a four-track recorder. The recordings can then be used to drive four vibration generators to carry out any further vehicle tests in the laboratory. Tests can thus be carried out independently of weather conditions and with considerably more instrumentation than would be possible in a normal field trial.

Increasing contact life of silver cadmium oxide contacts by up to 40% is possible by adding cobalt to them according to recent tests by Texas Instruments. Such silver cadmium oxide with additives is patented and will be marketed under the trade mark T1-CAD.